THE SWITCHMODE GUIDE

LINEAR AND POWER PRODUCTS

MOTOROLA

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The Designer's Guide for Switching Power Supply Circuits and Components

Introduction

The Switching Power Supply continues to increase in popularity and is one of the fastest growing markets in the world of power conversion. Its performance and size advantages meet the needs of today's modern and compact electronic equipments and the increasing variety of components directed at these applications makes new designs even more practical.

This guide is intended to provide the designer with an overview of the more popular inverter circuits, their basic theory of operation, and some of the subtle characteristics involved in selecting a circuit and the appropriate components. Also included are valuable design tips on both the major passive and active components needed for a successful design. Finally, a complete set of selector guides to Motorola's Switchmode components is provided which gives a detailed listing of the industry's most comprehensive line of semiconductor products for switching power supplies.

Comparison with Linear Regulators

The primary advantages of a switching power supply are efficiency, size, and weight. It is also a more complex design, cannot meet some of the performance capabilities of linear supplies and generates a considerable amount of electrical noise. Switchers are being accepted in the industry, particularly where size and efficiency are of prime importance, because its performance is still adequate for most applications and is often cost competitive in the 50 W power level and above. Because the switcher's passive components such as transformers and filters are smaller, they are almost always lower in cost than the high power (100 W) linear regulators. However, active component count is high (70 to 140 devices) and remains high regardless of the output power rating. This makes it less cost effective at the lower power levels. Switchers have been significantly cost reduced in recent years because designers have been able to simplify the control circuits with new, cost effective integrated circuits and have found even lower cost alternatives in the passive component area.

A performance comparison chart of switching versus linear supplies is shown in Figure 1. Switcher efficiencies run from 70 to 80% but occasionally fall to (60–65%) when linear post regulators are used for the auxillary outputs. Some linear power supplies on the other hand, are operated with up to 50% efficiency but these are areas where line variations or hold-up time problems are minimal. Most linears operate with typical efficiencies of only 30%. The overall size reduction of a 20 kHz switcher is about 4:1 and newer designs in the 100 to 200 kHz region end up at about 8:1 (versus a linear). Other characteristics such as static regulation specs are comparable, while ripple and load transient response are usually worse. Output noise specs can be somewhat misleading. Very often a 500 mV switching spike at the output may be attenuated considerably at the load itself due to the series inductance of the connecting cables and the additional filter capacitors found in many logic circuits. In the future, the noise generated at higher switching frequencies (100–500 kHz) will probably be easier to filter and the transient response will be faster. Hold-up time is the switchers inherent ability to regulate over wide variations in input voltage. It is easier to store energy in high voltage capacitors (200–400 V) than in the lower voltage (20–50 V) filter capacitors common to linear's power supplies. This is due to the fact that the physical size of a capacitor is dependent on its CV product while energy storage is proportional to CV^2.

Popular Inverter Configurations

A switching power supply is a relatively complex circuit as is shown by the four basic building blocks of Figure 2. It is apparent here that the heart of the supply is really the high frequency inverter. It is here that the work of chopping the rectified line at a high frequency (20 kHz) is done. It is here also that the line voltage is transformed down to the correct output level for use by logic or other electronic circuits. The remaining blocks support this basic function. The 60 Hz input line is rectified and filtered by one block and after the inverter steps this voltage down, the output is again rectified and filtered by another. The task of regulating the output voltage is left to the control circuit which closes the loop from the output to the inverter. Most control circuits generate a fixed frequency internally and utilize pulse width modulation techniques to implement the desired regulation. Basi-

FIGURE 1 — 20 kHz Switcher versus Linear Performance

Parameter	Switcher	Linear
Efficiency	75%	30%
Size	2.0 W/IN^3	0.5 W/IN^3
Weight	40 W/lb.	10 W/lb.
Cost 200–500 W*	$1.00/W	$1.25/W
Cost 50–150 W*	$1.50/W	$1.50/W
Line and Load Regulation	0.1%	0.1%
Output Ripple V_{P-P}	50 mV	5.0 mV
Noise V_{P-P}	50–200 mV	—
Transient Response	1 ms	20 µs
Hold-Up Time	20–30 ms	1–2 ms

*Based on 1980 Cost Figures

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cally, the on-time of the square wave drive to the inverter is controlled by the output voltage. As load is removed or input voltage increases, the slight rise in output voltage will signal the control circuit to deliver shorter pulses to the inverter and conversely as the load is increased or input voltage decreases, wider pulses will be fed to the inverter.

The inverter configurations used in today's switchers actually evolved from the buck and boost circuits shown in Figures 3A and 3B. In each case the regulating means and loop analysis will remain the same but a transformer is added in order to provide electrical isolation between the line and load. The forward converter family which includes the push-pull and half bridge circuits evolved

FIGURE 2 — Functional Block Diagram — Switching Power Supply

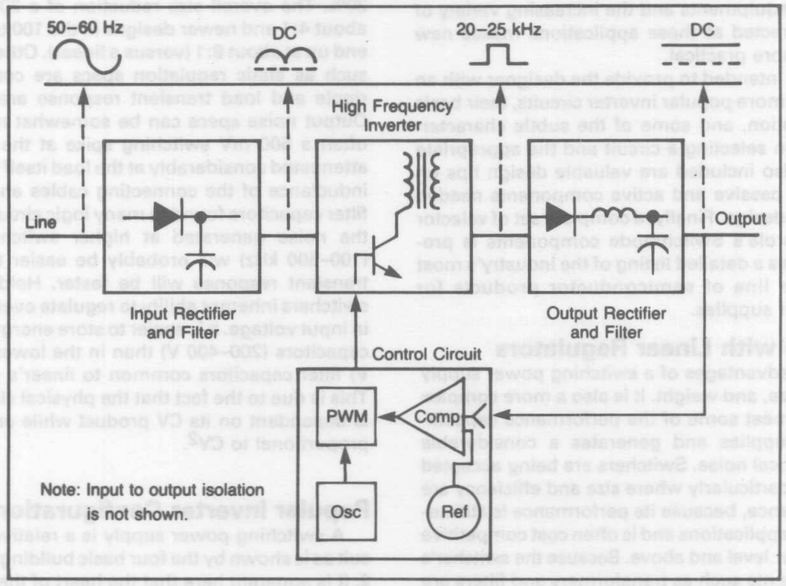
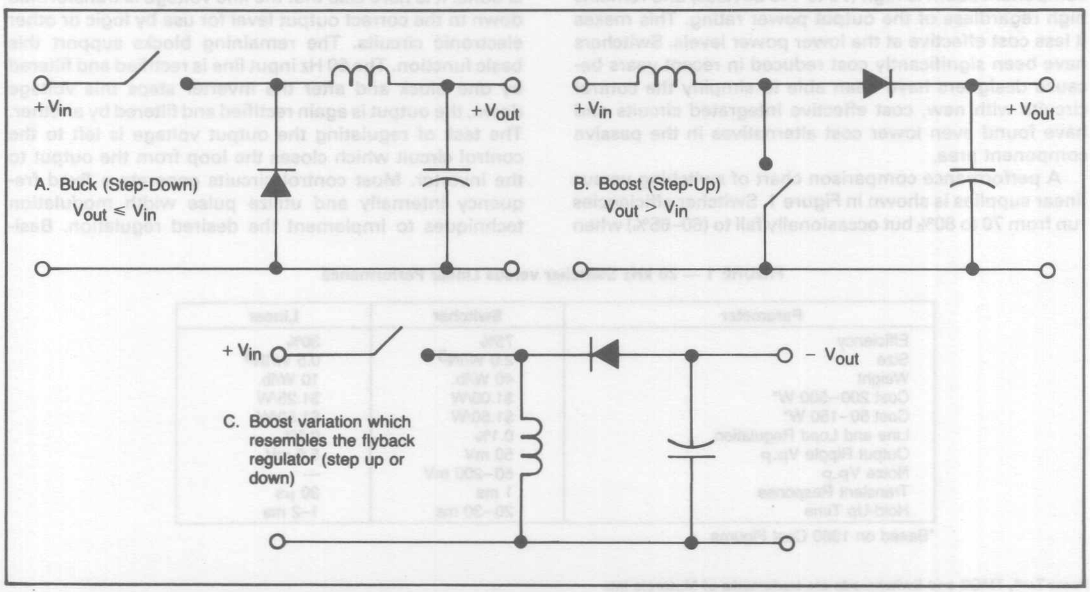


FIGURE 3 — Nonisolated DC-DC Converters



from the buck regulator (Figure 3A). And the newest switcher, the flyback converter, actually evolved from the boost regulator. The buck circuit interrupts the line and provides a variable pulse width square wave to a simple averaging LC filter. In this case, the first order approximation of the output voltage is $V_{out} = V_{in} \times$ duty cycle and regulation is accomplished by simply varying the duty cycle. This is satisfactory for most analysis work and only the transformer turns ration will have to be adjusted slightly to compensate for IR drops, diode drops, and transistor saturation voltages.

Operation of the boost circuit is more subtle in that it first stores energy in a choke and then additively delivers this energy with the input line to the load. However, the flyback regulators which evolved from this configuration delivers only the energy stored in the choke to the load. This method of operation is actually based on the boost variation model shown in Figure 3C. Here, when the switch is opened, only the stored inductive energy is delivered to the load. The true boost circuit can also regulate by stepping up (or boosting) the input voltage whereas the variation or flyback regulator can step the input voltage up or down. Analysis of the boost regulator begins by dealing with the choke as an energy storage element which delivers a fixed amount of power to the load:

$$P_o = 1/2\, L\, I^2 f_o$$

where I = the peak choke current
f_o = the operating frequency
and L = the inductance

Because it delivers a fixed amount of power to the load regardless of load impedance (except for short circuits), the boost regulator is the designer's first choice in photo-flash and capacitive-discharge (CD) automotive ignition circuits to recharge the capacitive load. It also makes a good battery charger. For an electronic circuit load, however, the load resistance must be known in order to determine the output voltage:

$$V_o = \sqrt{P_o R_L} = I\sqrt{\frac{L f_o R_L}{2}}$$

where R_L = The load resistance

In this case, the choke current is proportional to the on-time or duty cycle of the switch and regulation for fixed loads simply involves varying the duty cycle as before. However, the output also depends on the load which was not the case with buck regulators and results in a variation of loop gain with load.

For both regulators, transient response or responses to step changes in load are very difficult to analyze. They lead to what is termed a "load dump" problem. This requires that energy already stored in the choke or filter be provided with a place to go when load is abruptly removed. Practical solutions to this problem include limiting the minimum load and using the right amount of filter capacitance to give the regulator time to respond to this change.

Flyback and Forward Converters

To take advantage of the regulating techniques just discussed and also provide isolation, a total of seven popular configurations have evolved and are illustrated in Figures 4 and 7. Each circuit has a practical power range or capability associated with it as follows:

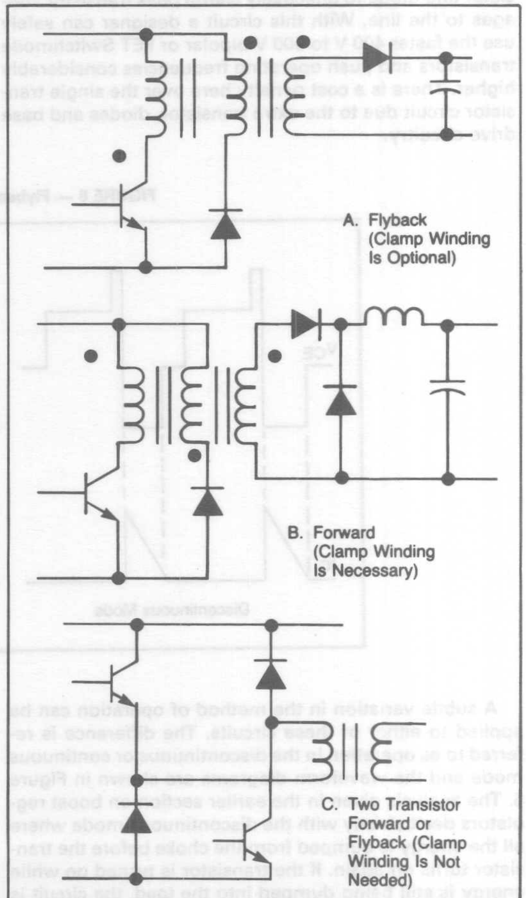
Circuit	Power Range	Motorola Reference
Flyback	50 to 100 watts	EB-87
Forward	100 to 200 watts	
Push-Pull	200 to 500 watts	EB-88, AN-737A
Half Bridge	200 to 500 watts	EB's 86 & 100, AN-767
Full Bridge	500 to 2000 watts	EB-85

First to be discussed will be the low power (20–200 W) converters which are dominated by the single transistor circuits shown in Figure 4. All of these circuits operate the magnetic element in the unipolar rather than bipolar mode. This means that transformer size is sacrificed for circuit simplicity.

Flyback

The flyback (alternately known as the "ringing choke") regulator stores energy in the primary winding and dumps it into the secondary windings (see Figure 4A). A clamp winding is usually present to allow energy stored

FIGURE 4 — Low Power Popular (20–200 W) Converter Configurations



in the leakage reactance to return safely to the line instead of avalanching the switching transistor. The operating model for this circuit is the boost circuit variation discussed earlier. The flyback is the lowest cost regulator (except at high power levels) because output filter chokes are not required since the output capacitors feed from an energy source rather than a voltage source. It does have higher output ripple than the forward converters because of this. However, it is an excellent choice when multiple output voltages are required and does tend to provide better cross regulation than the other types. In other words changing the load on one winding will have little effect on the output voltage of the others.

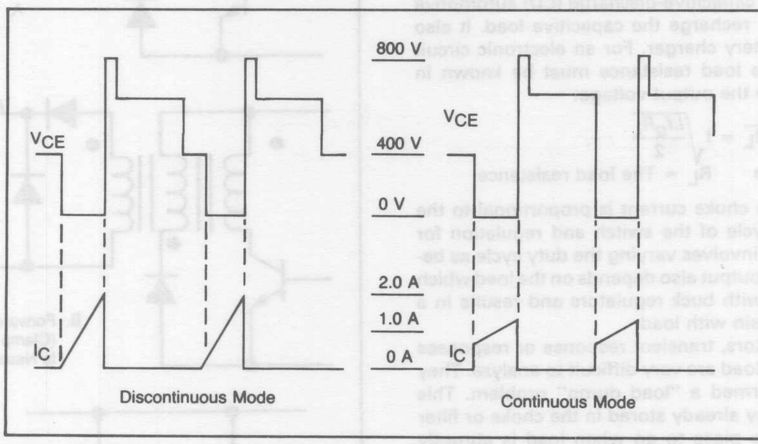
A 120/220 Vac flyback design requires transistors that block twice the peak line plus transients or about 1.0 kV. Presently variations of the 1200 to 1500 V horizontal deflection transistors are used here. These bipolar devices are relatively slow (t_f = 1.0 μs) and tend to limit efficient operating frequencies to 20 to 30 kHz. The availability of 1000 V TMOS FETs will permit operation at much higher frequencies. Faster 1.0 kV bipolar transistors are also planned in the future and will provide another design alternative. The two transistor variations of this circuit (Figure 4C) eliminates the clamp winding and adds a transistor and diode to effectively clamp peak transistor voltages to the line. With this circuit a designer can safely use the faster 400 V to 500 V bipolar or FET Switchmode transistors and push operating frequencies considerably higher. There is a cost penalty here over the single transistor circuit due to the extra transistor, diodes and base drive circuitry.

operating in the continuous mode. This is generally an advantage for the transistor in that it needs to switch only half as much peak current in order to deliver the same power to the load. In many instances, the same transformer may be used with only the gap reduced to provide more inductance. Sometimes the core size will need to be increased to support the higher LI product (2 to 4 times) now required because the inductance must increase by almost 10 times to effectively reduce the peak current by two. In dealing with the continuous mode, it should also be noted that the transistor must now turn on from 500 to 600 V rather than 400 V level because there no longer is any dead time to allow the flyback voltage to settle back down in the input voltage level. Generally it is advisable to have $V_{CEO\,(SUS)}$ ratings comparable to the turn-on requirements.

The flyback converter stands out from the others in its need for a low inductance, high current primary. Conventional E and pot core ferrites are difficult to work with because their permeability is too high even with relatively large gaps (50 to 100 milli-inches). The industry needs something better (like powered iron) that will provide permeabilities of 60 to 120 instead of 2000 to 3000 for this application.

The single transistor forward converter is shown in Figure 4B. Although it initially appears very similar to the flyback, it is not. The operating model for this circuit is actually the buck regulator discussed earlier. Instead of storing energy in the transformer and then delivering it to the load, this circuit uses the transformer in the active or forward mode and delivers power to the load while

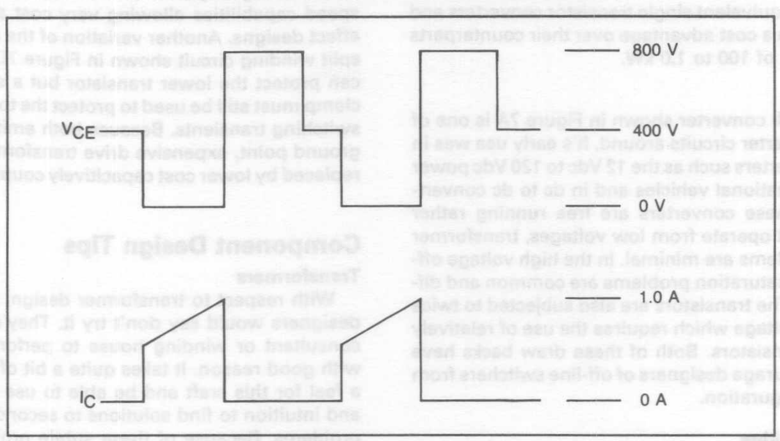
FIGURE 5 — Flyback Transistor Waveforms



A subtle variation in the method of operation can be applied to either of these circuits. The difference is referred to as operation in the discontinuous or continuous mode and the waveform diagrams are shown in Figure 5. The analysis given in the earlier section on boost regulators dealt strictly with the discontinuous mode where all the energy is dumped from the choke before the transistor turns on again. If the transistor is turned on while energy is still being dumped into the load, the circuit is

the transistor is on. The additional output rectifier is used as a freewheeling diode from the LC filter and the third winding is actually a reset winding. It generally has the same turns as the primary, (is usually bifilar wound) and does clamp the reset voltage to twice the line. However, its main function is to return energy stored in the magnetizing inductance to the line and thereby reset the core after each cycle of operation. Because it takes the same time to set and reset the core, the duty cycle of this circuit

FIGURE 6 — Forward Converter Transistor Waveforms

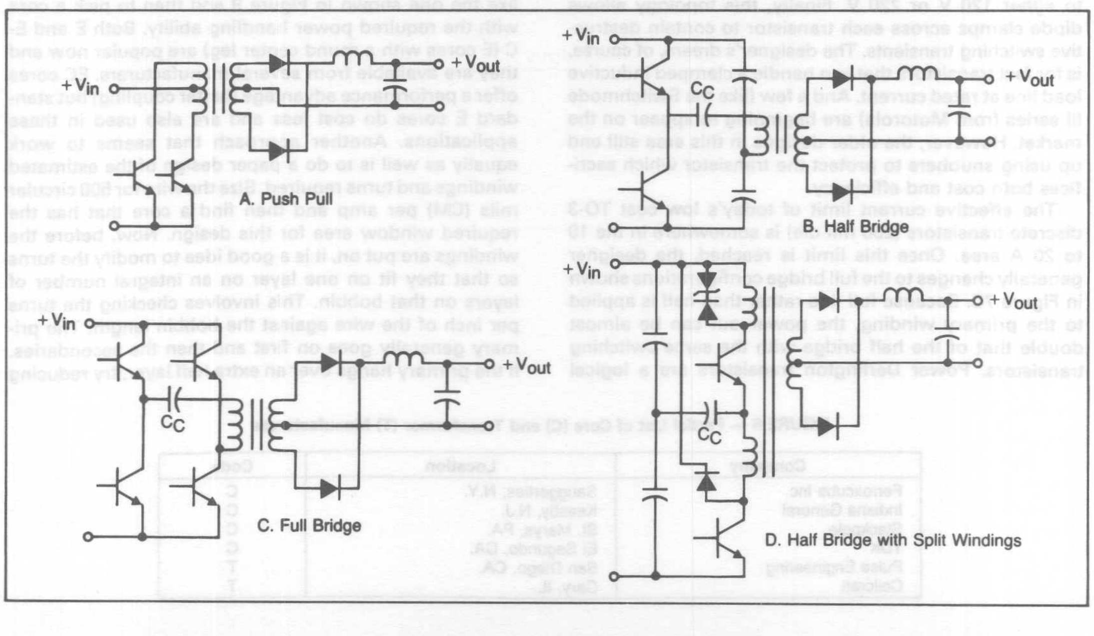


cannot exceed 50%. This also is a very popular low power converter and like the flyback is practically immune from transformer saturation problems. Transistor waveforms shown in Figure 6 illustrate that the voltage requirements are identical to the flyback. For the single transistor versions, 400 V turn-on and 1.0 kV blocking devices like the 1200 to 1500 V transistors are required. The two transistor circuit variations shown in Figure 4C again adds a cost penalty but allows a designer to use the faster 400 to 500 V devices. With this circuit, operation in the discontinuous mode refers to the time when the load is reduced to a point where the filter choke runs "dry". This means that choke current starts at and returns to zero during each cycle of operation. Most designers prefer to avoid this type of mode because of higher ripple and noise even though there are no adverse effects on the components themselves. Standard ferrite cores work fine here and in the high power converters as well. In these applications, no gap is used as the high permeability (3000) results in the desirable effect of very low magnetizing current levels.

Push-Pull and Bridge Converters

The high power circuits shown in Figure 7 all operate the magnetic element in the bipolar or push-pull mode and require 2 to 4 inverter transistors. Because the trans-

FIGURE 7 — High Power Popular Converter Configurations (100 W–1.0 kW)



formers operate in this mode they tend to be almost half the size of the equivalent single transistor converters and thereby provide a cost advantage over their counterparts at power levels of 100 to 1.0 kW.

Push-Pull

The push-pull converter shown in Figure 7A is one of the oldest converter circuits around. It's early use was in low voltage inverters such as the 12 Vdc to 120 Vdc power source for recreational vehicles and in dc to dc converters. Because these converters are free running rather than driven and operate from low voltages, transformer saturation problems are minimal. In the high voltage off-line switchers, saturation problems are common and difficult to solve. The transistors are also subjected to twice the peak line voltage which requires the use of relatively slow 1.0 kV transistors. Both of these draw backs have tended to discourage designers of off-line switchers from using this configuration.

Half and Full Bridge

The most popular high power converter is the half bridge (Figure 7B). It has two clear advantages over the push-pull and became the favorite rather quickly. First, the transistors never see more than the peak line voltage and the standard 400 V fast Switchmode transistors that are readily available may be used. And second, and probably even more important, transformer saturation problems are easily minimized by use of a small coupling capacitor (about 2 – 5 μF) as shown. Because the primary winding is driven in both directions, a full wave output filter, rather than half, is now used and the core is actually utilized more effectively. Another more subtle advantage of this circuit is that the input filter capacitors are placed in series across the rectified 220 V line which allows them to be used as the voltage doubler elements on a 120 V line. This still allows the inverter transformer to operate from a nominal 320 V bus when the circuit is connected to either 120 V or 220 V. Finally, this topology allows diode clamps across each transistor to contain destructive switching transients. The designer's dream, of course, is for fast transistors that can handle a clamped inductive load line at rated current. And a few (like the Switchmode III series from Motorola) are beginning to appear on the market. However, the older designs in this area still end up using snubbers to protect the transistor which sacrifices both cost and efficiency.

The effective current limit of today's low cost TO-3 discrete transistors (250 mil die) is somewhere in the 10 to 20 A area. Once this limit is reached, the designer generally changes to the full bridge configurations shown in Figure 7C. Because full line rather than half is applied to the primary winding, the power out can be almost double that of the half bridge with the same switching transistors. Power Darlington transistors are a logical choice to higher power control with current, voltage and speed capabilities allowing very cost and performance effect designs. Another variation of the half bridge is the split winding circuit shown in Figure 7D. A diode clamp can protect the lower transistor but a snubber or zener clamp must still be used to protect the top transistor from switching transients. Because both emitters are at an ac ground point, expensive drive transformers can now be replaced by lower cost capacitively coupled drive circuits.

Component Design Tips

Transformers

With respect to transformer design, many of today's designers would say don't try it. They'd advise using a consultant or winding house to perform this task and with good reason. It takes quite a bit of time to develop a feel for this craft and be able to use both experience and intuition to find solutions to second and third order problems. Because of these subtle problems, most designers find that after the first paper design is done, as many as four or five lab iterations may be necessary before the transformer meets the design goals. However, there is a considerable design challenge in this area and a great deal of satisfaction can be obtained by mastering it.

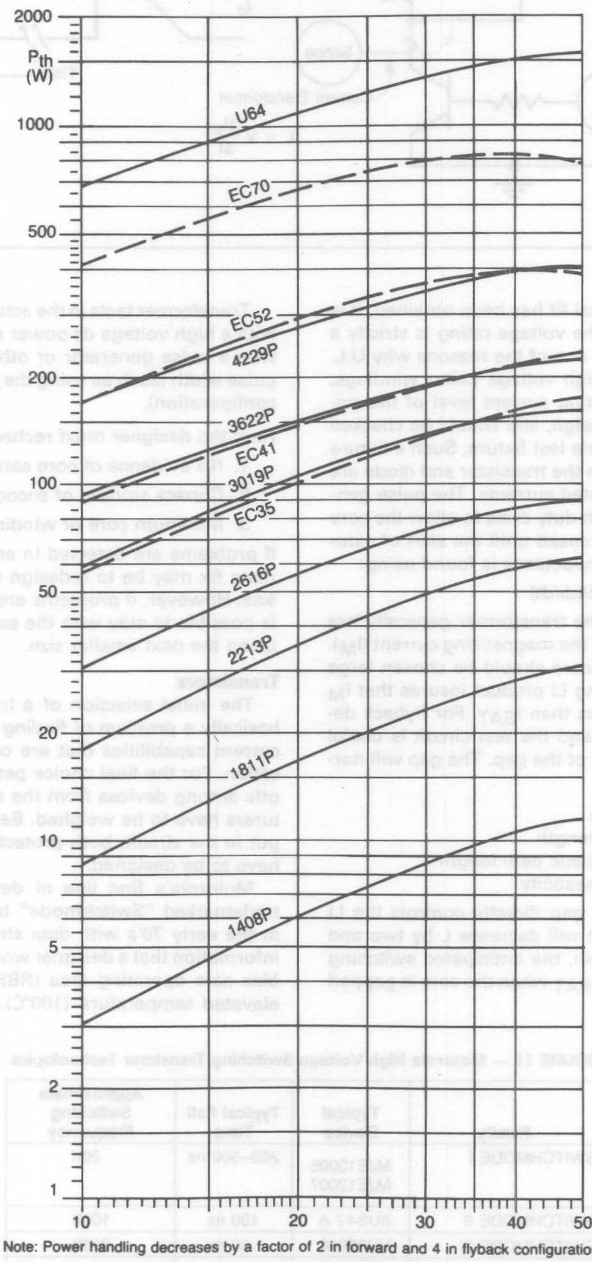
This component design, as does all others, begins by requesting all available literature from the appropriate manufacturers and then following this up with phone calls when specific questions arise. A partial list of companies is shown in Figure 8. Designs below 50 W generally use pot cores but for 50 W and above E cores are preferred. E cores expose the windings to air so that heat is not trapped inside and make it easier to bring out connections for several windings. Remember that flyback designs require lower permeability cores than the others. The classic approach is to consult manufacturers charts like the one shown in Figure 9 and then to pick a core with the required power handling ability. Both E and E-C (E cores with a round center leg) are popular now and they are available from several manufacturers. EC cores offer a performance advantage (better coupling) but standard E cores do cost less and are also used in these applications. Another approach that seems to work equally as well is to do a paper design of the estimated windings and turns required. Size the wire for 500 circular mils (CM) per amp and then find a core that has the required window area for this design. Now, before the windings are put on, it is a good idea to modify the turns so that they fit on one layer on an integral number of layers on that bobbin. This involves checking the turns per inch of the wire against the bobbin length. The primary generally goes on first and then the secondaries. If the primary hangs over an extra half layer, try reducing

FIGURE 8 — Partial List of Core (C) and Transformer (T) Manufacturers

Company	Location	Code
Ferroxcube Inc.	Saugerties, N.Y.	C
Indiana General	Keasby, N.J.	C
Stackpole	St. Marys, PA.	C
TDK	El Segundo, CA.	C
Pulse Engineering	San Diego, CA.	T
Coilcraft	Cary, IL.	T

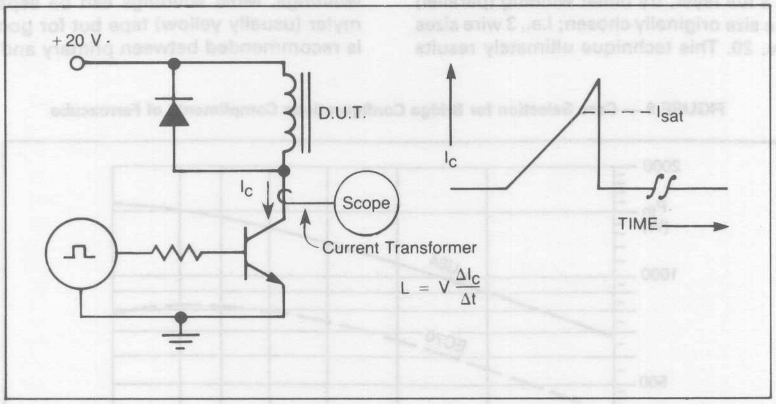
the turns or the wire size. Conversely, if the secondary does not take up a full layer, try bifilar winding (parallel) using wire half the size originally chosen; i.e., 3 wire sizes smaller like 23 vs. 20. This technique ultimately results in the use of foil for the higher current (20 A) low voltage windings. Most windings can be separated with 3 mil mylar (usually yellow) tape but for good isolation, cloth is recommended between primary and secondary.

FIGURE 9 — Core Selection for Bridge Configurations Compliments of Ferroxcube



Note: Power handling decreases by a factor of 2 in forward and 4 in flyback configurations.

FIGURE 10 — Simple Coil Tester



Finally, once a mechanical fit has been obtained, it is time for the circuit tests. The voltage rating is strictly a mechanical problem and is one of the reasons why U.L. normally does not allow high voltage bifilar windings. The inductance and saturating current level of the primary are inherent to the design, and should be checked in the circuit or other suitable test fixture. Such a fixture is shown in Figure 10 where the transistor and diode are sized to handle the anticipated currents. The pulse generator is run at a low enough duty cycle to allow the core to reset. Pulse width is increased until the start of saturation is observed (I_{SAT}). Inductance is found using:

$$L = E/(di/dt)$$

In forward converters, the transformer generally has no gap in order to minimize the magnetizing current (I_M). For these applications the core should be chosen large enough so that the resulting LI product insures that I_M at operating voltages is less than I_{SAT}. For flyback designs, a gap is necessary and the test circuit is useful again to evaluate the effect of the gap. The gap will normally be quite large

$$L_g >> L_m/u$$

where: L_g = gap length
L_m = magnetic path length
u = permeability

Under this stipulation, the gap directly controls the LI parameters and doubling it will decrease L by two and increase I_{SAT} by two. Again, the anticipated switching currents must be less than I_{SAT} when the core is gapped for the correct inductance.

Transformer tests in the actual supply are usually done with a high voltage dc power supply on the primary and with a pulse generator or other manual control for the pulse width (such as using the control IC in the open loop configuration).

Here the designer must recheck three areas:

1. No evidence of core saturation
2. Correct amount of secondary voltage
3. Minimum core or winding heat rise

If problems are detected in any of these areas, the ultimate fix may be to redesign using the next larger core size. However, if problems are minimal, or none exist, it is possible to stay with the same core or even consider using the next smaller size.

Transistors

The initial selection of a transistor for a switcher is basically a problem of finding the one with voltage and current capabilities that are compatible with the application. For the final choice performance and cost tradeoffs among devices from the same or several manufacturers have to be weighed. Before these devices can be put in the circuit, both protective and drive circuits will have to be designed.

Motorola's first line of devices for switchers were trademarked "Switchmode" transistors and introduced in the early 70's with data sheets that provided all the information that a designer would need including reverse bias safe operating area (RBSOA) and performance at elevated temperature (100°C). The first series was the

FIGURE 11 — Motorola High Voltage Switching Transistor Technologies

Family	Typical Device	Typical Fall Time	Approximate Switching Frequency
SWITCHMODE I	MJE13005 MJE12007	200–500 ns	20K
SWITCHMODE II	BUS47 A	100 ns	100K
SWITCHMODE III	MJ16010	50 ns	200K
TMOS	MTP5N40	20 ns	500K

2N6542 through 2N6547, TO-3 and was followed by the MJE13002 through MJE13009 series in a plastic TO-220 package. Finally, high voltage (1.0 kV) requirements were met by the metal MJ8500 thru MJ8505 series and the plastic MJE8500 series. And just recently, Motorola introduced the two new families of "Switchmode" transistors shown in Figure 11. The Switchmode II series is an advanced version of Switchmode I that features faster switching. Switchmode III is the state of today's bipolar art with both exceptional speed and RBSOA. Here, device cost is somewhat higher, but system costs may be lowered because of reduced snubber requirements and higher operating frequencies. A similar argument applies to Motorola TMOS Power FETs. These devices make it possible to switch efficiently at higher frequencies (200 to 500 kHz) but the main selling point is that they are easier to drive. This latter point is the one most often made to show that systems savings are again quite possible even though the initial device cost is higher.

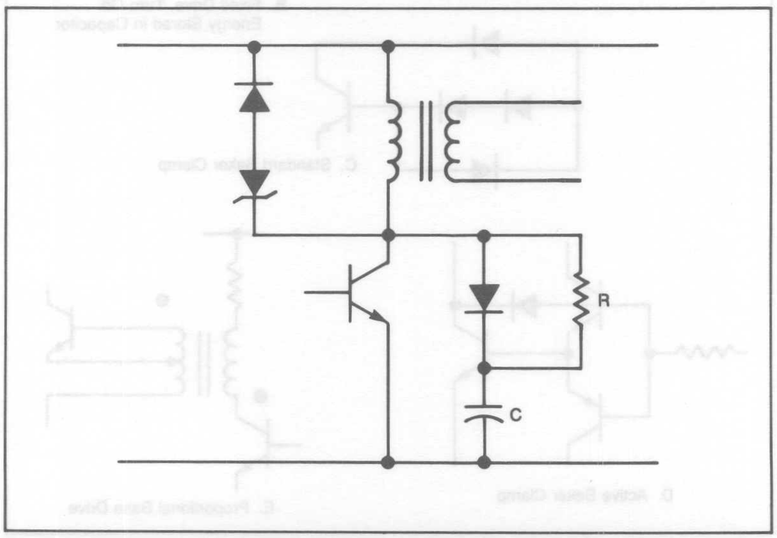
Most Switchmode transistor load lines are inductive during turn-on and turn-off. Turn-on is generally inductive because the short circuit created by output rectifier reverse recovery times is isolated by leakage inductance in the transformer. This inductance effectively snubs most turn-on load lines so that the rectifier recovery (or short circuit) current and the input voltage are not applied simultaneously to the transistor. Sometimes primary interwinding capacitance presents a small current spike but usually turn-on transients are not a problem. Turn-off transients due to this same leakage inductance, however, are almost always a problem. In bridge circuits, clamp diodes can be used to limit these voltage spikes. If the resulting inductive load line exceeds the transistors reverse bias switching capability (RBSOA) then an RC network may also be added across the primary to absorb some of this transient energy. The time constant of this network should equal the anticipated switching time of the transistor (100 ns to 1.0 μs). Resistance values of 100 to 1000 ohms in this RC network are generally appropriate. Trial and error will indicate how low the resistor has to be to provide the correct amount of snubbing. For single transistor converters, the snubber shown in Figure 13 is generally used. Here slightly different criterion are used to define the R and C values:

FIGURE 12 — Power Transistor Voltage Chart

Line Voltage	Circuit: Flyback, Forward or Push-Pull		Circuit: Half or Full Bridge	
	V_{CEV}	$V_{CEO(sus)}$	$V_{CEO(sus)}$	V_{CEV}
220	850	400	400	400
120	450	200	200	200

Figure 12 is a review of the transistor voltage requirements for the various off-line converter circuits. As illustrated, the most stringent requirement for single transistor circuits (flyback and forward) is the blocking or V_{CEV} rating. Bridge circuits, on the other hand, turn-on and off from the dc bus and their most critical voltage is the turn-on or $V_{CEO\ (SUS)}$ rating.

FIGURE 13 — Zener Clamp and Snubber for Single Transistor Converters



$$C = \frac{I\,t_f}{V}$$

where I = The peak switching current
t_f = The transistor fall time
V = The peak switching voltage (Approximately twice the dc bus)

also $R = t_{on}/C$ (it is not necessary to completely discharge this capacitor in order to obtain the desired effects of this circuit)

where t_{on} = The minimum on-time or pulse width

and $P_R = \frac{CV^2 f}{2}$

where P_R = The power rating of the resistor
and f = The operating frequency

Most of today's transistors that are used in 20 kHz converters switch slow enough so that most of the energy stored in the leakage inductance is dissipated by the snubber or transistor causing little voltage overshoot. Higher speed converters and transistors present a slightly different problem. In these newer designs snubber elements are smaller and voltage spikes from energy left in the leakage inductance may be a more critical problem depending on how good the coupling is between the primary and clamp windings.

Zener Diodes

If necessary, protection from voltage spikes may be obtained by adding a zener and rectifier across the primary as shown in Figure 13. Here Motorola's 1 W and 5 W zener lines with ratings up to 200 V can provide the clamping or spike limiting function. If the zener must handle most of the power, its size can be estimated using:

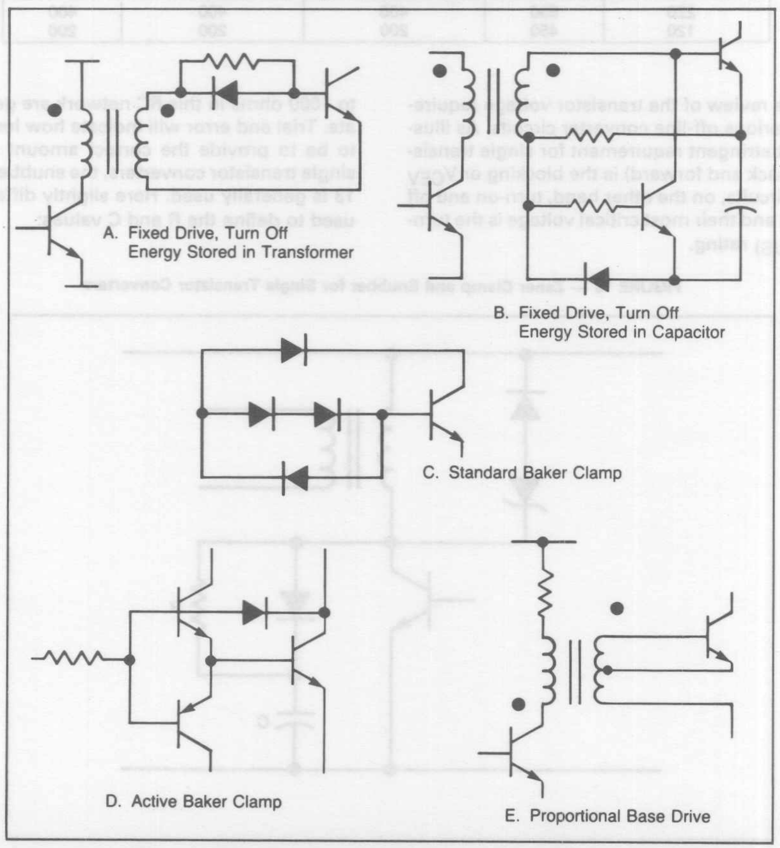
$$P_Z = \frac{L_L\, I^2 f}{2}$$

where P_Z = The zener power rating
and L_L = The leakage inductance (measured with the clamp winding or secondary shorted)
I = Peak collector current
f = Operating frequency

Mosorb Transient Suppressors

Distinction is sometimes made between devices trademarked Mosorb (by Motorola Inc.), and standard zener/avalanche diodes used for reference, low-level regulation and low-level protection purposes. It must be emphasized that Mosorb devices are, in fact, zener diodes. The

FIGURE 14 — Typical Bipolar Base Drive Circuits



basic semiconductor technology and processing are identical. The primary difference is in the applications for which they are designed. Mosorb devices are intended specifically for transient protection purposes and are designed, therefore, with a large effective junction area that provides high pulse power capability while minimizing the total silicon use. Thus, Mosorb pulse power ratings begin at 600 watts — well in excess of low power conventional zener diodes which in many cases do not even include pulse power ratings among their specifications.

MOVs, like Mosorbs, do have the pulse power capabilities for transient suppression. They are metal oxide varistors (not semiconductors) that exhibit bidirectional avalanche characteristics, similar to those of back-to-back connected zeners. The main attributes of such devices are low manufacturing cost, the ability to absorb high energy surges (up to 600 joules) and symmetrical bidirectional "breakdown" characteristics. Major disadvantages are: high clamping factor, an internal wear-out mechanism and an absence of low-end voltage capability. These limitations restrict the use of MOVs primarily to the protection of insensitive electronic components against high energy transients in applications above 20 volts, whereas, Mosorbs are best suited for precise protection of sensitive equipment even in the low voltage range — the same range covered by conventional zener diodes.

Drive Considerations

There are probably as many base drive circuits for bipolars as there are designers. Ideally, the transistor would like just enough forward drive (current) to stay in or near saturation and reverse drive that varies with the amount of stored base charge such as a low impedance reverse voltage. Many of today's common drive circuits are shown in Figure 14. The fixed drive circuits of 14A and 14B tend to emphasize economy, while the Baker clamp and proportional drive circuits of 14C, 14D and 14E emphasize performance over cost.

FET drive circuits are another alternative. The standard that has evolved at this time is shown in Figure 15. This transformer coupled circuit will produce forward and reverse voltages applied to the FET gate which vary with the duty cycle as shown. For this example, a V_{GS} rating of 20 V would be adequate for the worst case condition of high logic supply (12 V) and minimum duty cycle. And yet, minimum gate drive levels of 10 V are still available with duty cycles up to 50%. If wide variations in duty cycle are anticipated, it might be wise to consider using a semi-regulated logic supply for these situations. Finally, one point that is not obvious when looking at the circuit is that FETs can be directly coupled to many ICs with only 100 mA of sink and source capability and still switch efficiently at 20 kHz. However, to achieve switching efficiently at higher frequencies, several amps of drive may be required on a pulsed basis in order to quickly charge and discharge the gate capacitances. A simple example will serve to illustrate this point and also show that the Miller effect, produced by C_{DG}, is the predominant speed limitation when switching high voltages (see Figure 15B). A FET responds instantaneously to changes in gate voltage and will begin to conduct when the threshold is reached (V_{GS} = 2 to 3 V) and be fully on with V_{GS} = 7 to 8 V. Gate waveforms will show a step at a point just above the threshold voltage which varies in duration depending on the amount of drive current available which determines both the rise and fall times for the drain current. To estimate drive current requirements, two simple calculations with gate capacitances can be made:

1. $I_M = C_{DG}dv/dt$ and
and 2. $I_G = C_{GS}dv/dt$

FIGURE 15A — Typical Transformer Coupled FET Drive

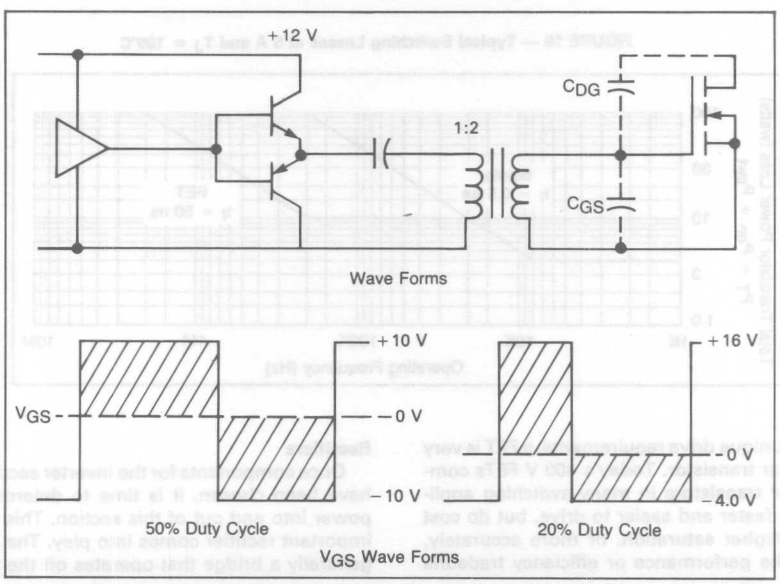
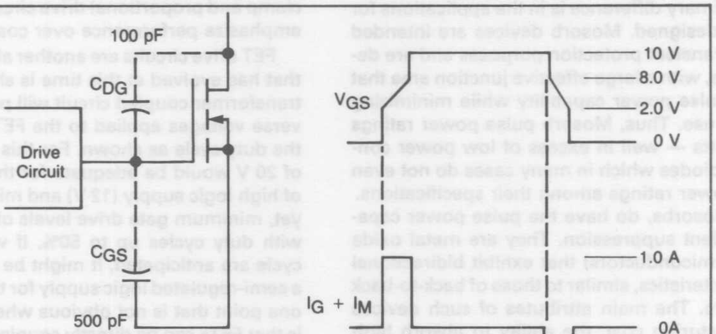


FIGURE 15B — FET Drive Current Requirements



1. Miller Current for 30 ns

$I_M = C_{DG}\ dv/dt$

$= 100\ pF \times \dfrac{300\ V}{30\ ns} = 1.0\ A$

2. Gate Cap Current for 30 ns

$I_G = C_{GS}\ dv/dt$

$= 500\ pF \times \dfrac{6.0\ V}{30\ ns} = 0.1\ A$

I_M is the current required by the Miller effect to charge the drain-to-gate capacitance at the rate it is desired to move the drain voltage (and current). And I_G is usually the lesser amount of current required to charge the gate-to-source capacitance through the linear region (2 to 8 V). As an example, if 30 ns switching times are desired at 300 V where C_{DG} = 100 pF and C_{GS} = 500 pF, then

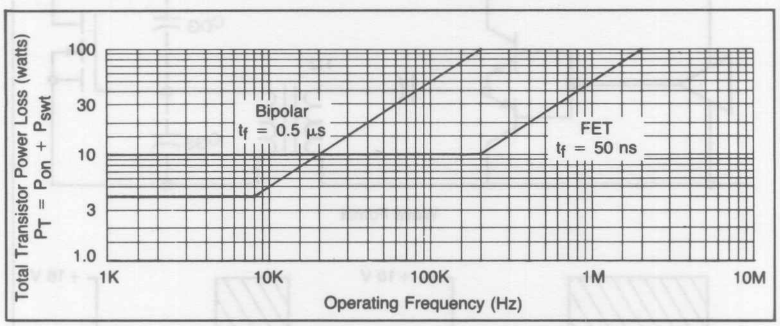
I_M = 100 pF x 300 V/30 ns 1.0 A and

I_G = 500 pF x 6 V/30 ns = 0.1 A

This example shows the direct proportion of drive current capability to speed and also illustrates that for most devices, C_{DG} will have the greatest effect on switching speed and that C_{GS} is important only in estimating turn-on and turn-off delays.

are analyzed using Figure 16. Here, typical power losses for 5 A switching transistors versus frequency are shown. The FET (and bipolar) losses were calculated at 100°C rather than 25°C because on resistance and switching times are highest here and 100°C is typical of many applications. These curves are asymptotes of the actual device performance, but are useful in establishing the "break point" of various devices, which is the point where saturation and switching losses are equal. Since this is as low as 10K for some bipolars, it is possible that a FET even with high on-voltages can be competitive efficiency-wise at 200 kHz. The faster Switchmode II and III bipolar products would fall somewhere between the curves shown and therefore, be more competitive with FETs at the higher operating frequencies.

FIGURE 16 — Typical Switching Losses at 5 A and T_J = 100°C



Aside from its unique drive requirements, a FET is very similar to a bipolar transistor. Today's 400 V FETs complete with bipolar transistors in many switching applications. They are faster and easier to drive, but do cost more and have higher saturation, or more accurately, "on" voltages. The performance or efficiency tradeoffs

Rectifiers

Once components for the inverter section of a switcher have been chosen, it is time to determine how to get power into and out of this section. This is where the all important rectifier comes into play. The input rectifier is generally a bridge that operates off the ac line and into

a capacitive filter. For the output section, most designers use Schottkys for efficient rectification of the low voltage, 5 V output windings and for the higher voltage, 12 to 15 V outputs, the more economical fast recovery or ultrafast diodes are used.

4. If $I_S < I_p$, consider either increasing the limiting resistor (R_S) or utilizing a larger diode.

In the output section where high frequency rectifiers are needed, there are several types available to the de-

FIGURE 17 — Choosing Input Rectifiers

	SBR	UFR	FR
V_F	0.5–0.6	0.9–1.0	1.2–1.4
t_{rr}	10 ns	25 ns	150 ns
t_{rr} FORM	"SOFT"	"ABRUPT"	"EITHER"
V_R	30–50 V	50–150 V	50–600 V

NOTES: 1. Low V_F improves efficiency
2. Low t_{rr} reduces transistor switching losses
3. Soft (verses abrupt) recovery reduces noise

For the process of choosing an input rectifier, it is useful to visualize the circuit shown in Figure 18. To reduce cost, most earlier approaches of using choke input filters, soft start relays (Triacs), or SCRs to bypass a large limiting resistor have been abandoned in favor of using small limiting resistors or thermistors and a large bridge. The bridge must be able to withstand the surge currents that exist from repetitive starts at peak line. The procedure for finding the right component and checking its fit is as follows:

1. Choose a rectifier with 2 to 5 times the average I_O required

2. Estimate the peak surge current (I_P) and time (t) using:

$$I_p = \frac{1.4\ V_{in}}{R_S} \qquad t = R_S C$$

Where V_{in} is the RMS input voltage
R_S is the total limiting resistance, and
C is the filter capacitor size

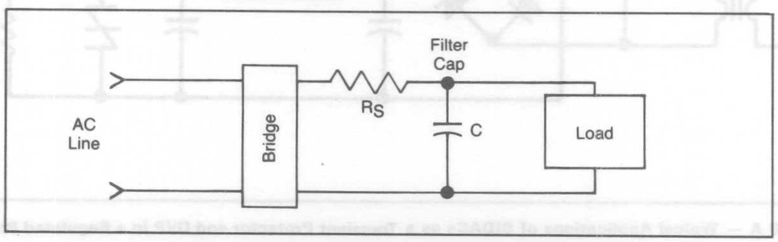
3. Compare this current pulse to the sub cycle surge current rating (I_S) of the diode itself. If the curve of I_S versus time is not given on the data sheet, the approximate value for I_S at a particular pulse width (t) may be calculated knowing:
   - I_{FSM} – the single cycle (8.3 ms) surge current rating and using.
   - $I^2 \sqrt{t} = K$ which applies when the diode temperature rise is controlled by its thermal response as well as power (i.e., $T = K'P \sqrt{t}$ for $t < 8$ ms. This gives:

$$I_S{}^2 \sqrt{t} = I^2{}_{FSM} \sqrt{8.3\ \text{ms}} \text{ or}$$

$$I_S = I_{FSM} \left(\frac{8.3\ \text{ms}}{t}\right)^{1/4}, \text{ t is in milliseconds}$$

signer. In addition to the Schottky (SBR) and fast recovery (FR), there is also an ultra fast recovery (UFR) which fills the gap between the 50 V Schottky and the 600 V fast recovery lines. Comparative performance for devices with similar current ratings is shown in Figure 18. The obvious point here is that lower forward voltage improves efficiency and lower recovery times reduces turn losses in the switching transistors, but the tradeoff is higher cost. As stated earlier, Schottkys are generally used for 5.0 V outputs and fast recovery devices for 12 V outputs and greater. The ultra fast is competing primarily with the Schottky in those applications where cost is more important than efficiency. Of these devices, only the Schottky may need special handling. Ten years ago Schottkys were very fragile and could fail short from either excessive dv/dt (1.0 to 5.0 volts per nanosecond) or reverse avalanche. Present day devices, however, have something similar to Motorola's "guard ring" and internal zener which minimizes these earlier problems and reduces the need for RC snubbers and other external protective networks.

FIGURE 18 — Output Rectifier Type Comparisons



SIDAC as a simple protection

The simple SIDAC circuit can also supply switchable load current. However, the conduction angle is not readily controllable, being a function of the peak applied voltage and the breakover voltage of the SIDAC. As an example, for peak line voltage of about 170 V, at $V_{(BO)}$ of 115 V and a holding current of 100 mA, the conduction angle would be about 130°. With higher peak input voltage (or lower breakdown voltages) the conduction angle would correspondingly increase. For non-critical conduction angle, 1.0 A rms switching applications, the SIDAC is a very cost-effective device.

Since the MK1V series of SIDACs have relatively tight $V_{(BO)}$ tolerances (104 V to 115 V for the – 115 device), other possible applications are over-voltage protection (OVP) and detection circuits. An example of this, as illustrated in Figure, is the SIDAC as a transient protector in the transformer-secondary of the medium voltage power supply, replacing the two more expensive back-to-back zeners or an MOV. The device can also be used across the output of the regulator (< 100 V) as a simple OVP but for this application, the regulator must have current foldback or a circuit breaker (or fuse) to minimize the dissipation of the SIDAC.

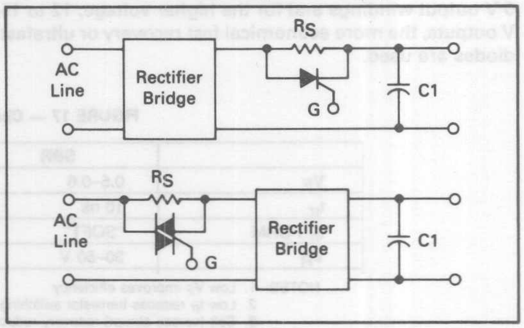
Power Triacs and Inrush Current Limiting

Many high current PWM switching supplies operate directly off the ac line. They have very large capacitive input filters with high inrush surge currents. The line circuit breaker and the rectifier bridge must be protected during turn-on.

Surge current limiting can be accomplished by adding R_S and an SCR "short" after charging C1 as shown in Figure 19, or by phase controlling the line voltage with a Triac.

For further information, see EB-78 and MC3420 data sheet.

FIGURE 19 — Surge Current Limiting For A Switching Power Supply

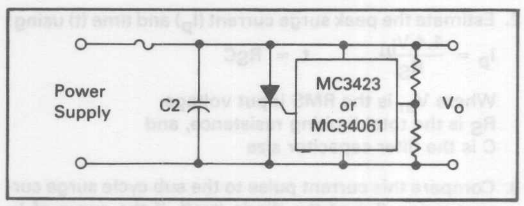


Power SCRs for Crowbar Applications

Linear and switching power supplies can be protected from overvoltage with a crowbar circuit. For linear supplies, the pass transistor can fail shorted, allowing high line transformer voltage to the load. For switching power supplies, a loose or disconnected remote sense lead can allow high voltage to the load.

The crowbar circuit, shown below ignores noise spikes but will fire the SCR when a valid overvoltage condition is detected. The SCR will discharge C2 and either blow the fuse or cause the power supply to shut down.

FIGURE 20 — Crowbar Circuit



For further information, see AN-789 and MC3423 data sheet.

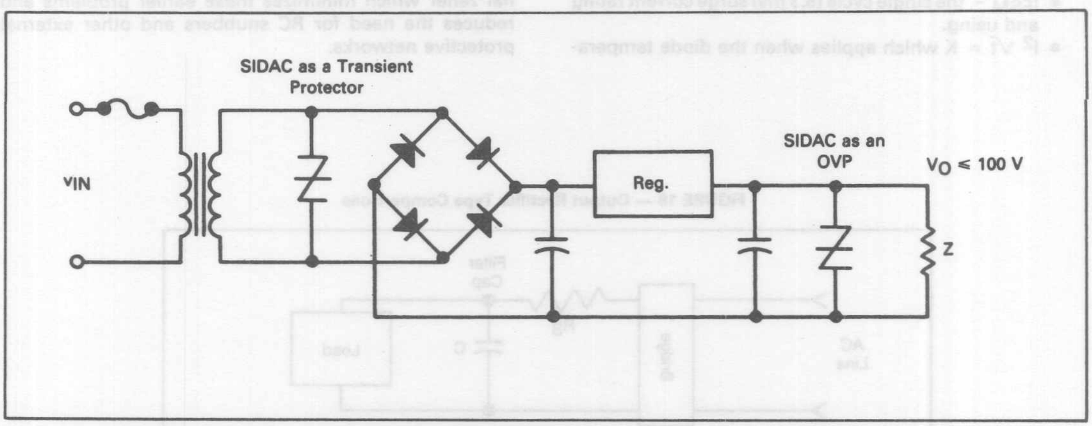


FIGURE 20 A — Typical Applications of SIDACs as a Transient Protector and OVP in a Regulated Power Supply

Capacitors and Filters

In today's 20 kHz switchers, aluminum electrolytics still predominate. The good news is that most have been characterized, improved, and cost reduced for this application. The input filter requires a voltage rating that depends on the peak line voltage; i.e., 400 to 450 V for a 220 V switcher. If voltage is increased beyond this point, the capacitor will begin to act like a zener and be thermally destroyed from high leakage currents if the rating is exceeded for enough time. In doubler circuits, voltage sharing of the two capacitors in series can be a problem. Here extra voltage capability may be needed to make up for the imbalances caused by different values of capacitance and leakage current. A bleeder resistor is normally used here not only for safety but to mask the differences in leakage current. The RMS current rating is also an important consideration for input capacitors and is an example of improvements offered by today's manufacturers. Earlier "lytics" usually lacked this rating and often overheated. Large capacitors that were not needed for performance were used just to reduce this heating. However, today's devices like the swedged variety from Mepco-Electra offer lower thermal resistance, improved connection to the foil and good RMS ratings. A partial list of manufacturers that supply both high voltage input and the lower voltage output capacitors for switchers is shown in Figure 21. Most of the companies offer not only the standard 85°C components, but devices with up to 125°C ratings which are required because of the high ambient temperatures (55 to 85°C) that many switchers have to operate in, many times without the benefit of fans.

For output capacitors the buzz word is low ESR (equivalent series resistance). It turns out that for most capacitors even in the so-called "low ESR" series, the output ripple depends more on this resistance than on the capacitor value itself. Although typical and maximum ESR ratings are now available on most capacitor designed for switchers, the lead inductance generally is not specified except for the ultra-high frequency four terminal capacitors from some vendors. This parameter is responsible for the relatively high switching spikes that appear at the output. However, at this point in time, most designers find it less costly and more effective to add a high frequency noise filter rather than use a relatively expensive capacitor with low equivalent series inductance (ESL).

These LC noise or spike filters are made using small powdered iron toriods (1/2 to 1" OD) with distributed windings to minimize interwinding capacitance. And the output is bypassed using a small 0.1 μF ceramic or a 10 to 50 μF tantalum or both. Larger powered iron toriods are often used in the main LC output filter although the higher permeability ferrite C and E cores with relatively large gaps can also be used. Calculations for the size of this component should take into account the minimum load so that the choke will not run "dry" as stated earlier.

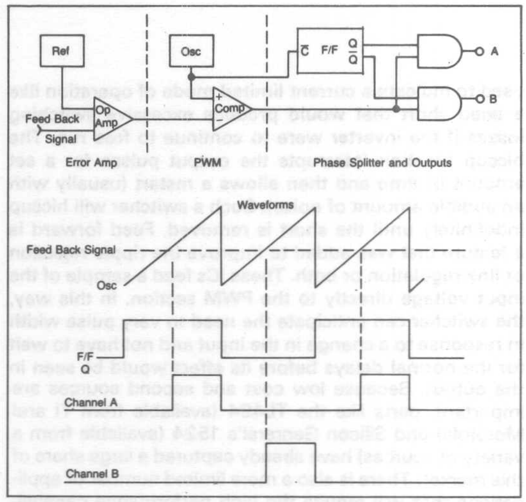
FIGURE 21 — Partial List of Capacitor Companies

Company (U.S.)	Location
Sprague	North Adams, MA
MEPCO/Electra	Columbia, SC
Cornell-Dublier	Sanford, NC
Sangamo	Pickens, SC
Mallory	Indianapolis, IN

Control Circuits

Ten years ago, discrete control circuits were in use and only bits and pieces of ICs could be found. Since that time, various semiconductor companies recognized the designers' needs for a dedicated control IC and now a variety of these circuits are on the market and widely used. They may provide the designer with a cost incentive over the discrete or a simpler control circuit or both. Internally, most of these resemble the functional configuration shown in Figure 22. The basic regulating function is performed in the pulse width modulator (PWM) section. Here, the dc feedback signal is compared to a fixed frequency sawtooth (or triangular) waveform. The result

FIGURE 22 — Basic SM Control IC



is a variable duty cycle pulse train which, with suitable buffer or interface circuits can be used to drive the power switching transistor. Some ICs provide only a single output while others provide the phase splitter shown to alternately pulse two output channels. In this latter case provisions are usually made either internally or by wire "or" ing the outputs to convert the dual to a single output channel. Additionally most ICs provide the error amplifier section shown as a means to process, compare and amplify the feedback signal.

Features required by a control IC vary to some extent because of the particular needs of a designer and on the circuit configuration chosen. However, most of today's current generation ICs have evolved with the capabilities or features listed in Figure 23. It is primarily the cost differences in these parts that determines whether all or only part of these features will be incorporated. Most of these are evident to the designer who has already started comparing data sheets, except perhaps for the hiccup and feed forward features. The hiccup terminology is

FIGURE 23 — Desirable Features of Switchmode Control ICs

- PROGRAMMABLE (TO 500 kHz) FIXED FREQUENCY OSCILLATOR
- LINEAR PWM SECTION WITH DUTY CYCLE FROM 0 TO 100 %
- ON BOARD ERROR AMPLIFIERS
- ON BOARD REFERENCE REGULATOR
- ADJUSTABLE DEAD TIME
- UNDERVOLTAGE (LOW V_{CC}) INHIBIT
- GOOD OUTPUT DRIVE (100 TO 200 mA)
- OPTION OF SINGLE OR DUAL CHANNEL OUTPUT
- UN-COMMITTED OUTPUT COLLECTOR AND EMITTER OR TOTEM POLE DRIVE CONFIGURATION
- SOFT START
- CURRENT LIMITING WITH "HICCUP MODE" AS BACKUP
- SYNC CAPABILITY

used to indicate a current limited mode of operation like a dead short that would produce excessive switching losses if the inverter were to continue to free run. The hiccup function interrupts the output pulses for a set amount of time and then allows a restart (usually with an audible amount of noise). Such a switcher will hiccup indefinitely until the short is removed. Feed forward is a feature that was added to improve the ripple rejection or line regulation or both. These ICs feed a sample of the input voltage directly to the PWM section. In this way, the switcher can anticipate the need to vary pulse width in response to a change in the input and not have to wait for the normal delays before its effect would be seen in the output. Because low cost and second sources are important, parts like the TL494 (available from TI and Motorola) and Silicon Genreral's 1524 (available from a variety of sources) have already captured a large share of this market. There is also a more limited number of applications that will require the high performance capabilities and will pay the cost premiums for parts like the SG 1526.

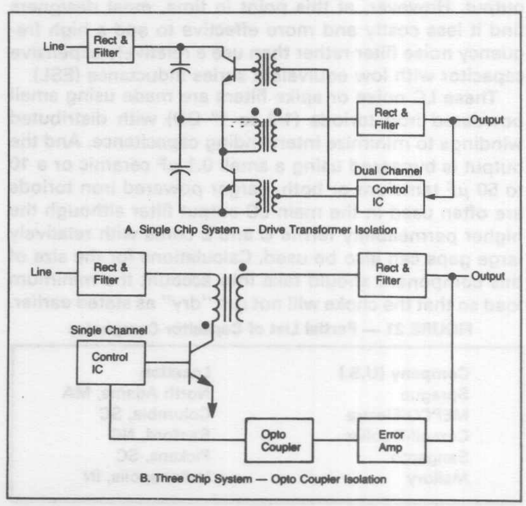
Today there is a need for a simple, low cost, single channel control IC for low power (20 to 200 W) applications like Motorola's MC34060 ans MC34063. This component would be used to run the low power flyback type configurations and probably would be part of a three chip rather than single chip system. The differences in these two approaches are illustrated in Figure 24.

When it is necessary to drive two or more power transistors, drive transformers are a practical interface element and are driven by the conventional dual channel IC just discussed (Figure 24A). In the case of a single transistor converter, howeyer, il is usually more cost effective to directly drive the transistor from the IC (Figure 24B). In this situation, an optocoupler is commonly used to couple the feedback signal from the output back to this control IC. And the error amplifier in this case is nothing more than an op amp and reference.

The Future

The future offers a lot of growth potential for switchers in general and low power switchers (50 – 200 watts) in particular. The latter are responding to the growth in microprocessor based equipment as well as computer peripherals. Today's configurations have already been challenged by the sine wave inverter which reduces noise and improves transistor reliability but does effect a cost penalty. Also, a trend to higher switching frequencies to reduce size and cost even further has begun. The latest bipolar can operate efficiently up to 100 kHz and the FET seems destined to own the 200 to 500 kHz range. These newer switchers have not yet realized a significant cost savings primarily because of deficiencies in the passive component area.

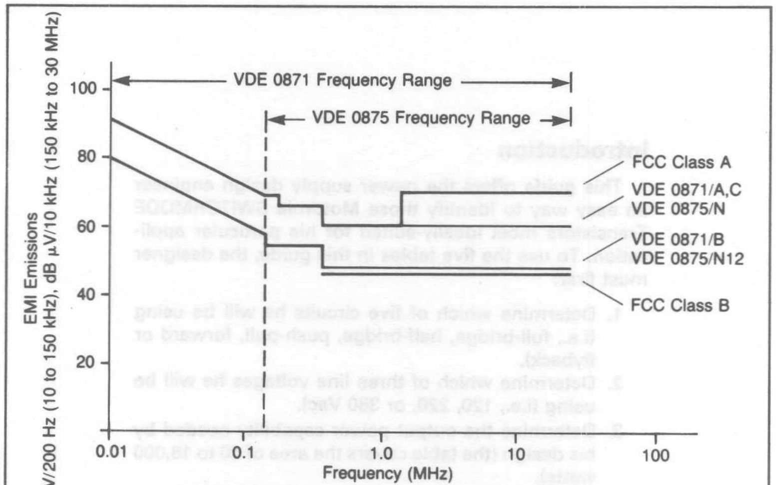
FIGURE 24 — Control Circuit Topologies



The growth pattern predicted at this time can possibly be impacted by noise problems. Originally governed only by MIL specs and the VDE in Europe, now (effective October 1981) the FCC has released a set of specifications that apply to electronic systems which often include switchers (see FCC Class A in Figure 25). It seems probable, however, that system engineers or power supply designers will be able to add the necessary line filters and EMI shields without evoking a significant cost penalty which would slow the growth of switchers.

The most optimistic note concerning switchers is in the component area. Switching power supply components have actually evolved from components used in similar applications. And it is very likely that newer and more mature products specifically for switchers will continue to appear over the next several years. The ultimate effect of this evolution will be to further simplify, cost reduce and increase the reliability of these designs.

FIGURE 25 — Noise Limits



Notes: 1. FCC Class A covers commercial, Class B covers residential.
2. Also SCE VDE 0871/0875 for noise and VDE 0730 or UL478 for safety.

Power Transistors

- **Basic Switching Power Supply Configurations**
- **Bipolar and TMOS Power MOSFETs**
- **Bipolar Discrete and Darlingtons**
- **TMOS Power MOSFETs**

Introduction

This guide offers the power supply design engineer an easy way to identify those Motorola SWITCHMODE Transistors most ideally-suited for his particular application. To use the five tables in this guide, the designer must first:

1. Determine which of five circuits he will be using (i.e., full-bridge, half-bridge, push-pull, forward or flyback).
2. Determine which of three line voltages he will be using (i.e., 120, 220, or 380 Vac).
3. Determine the output power capability needed by his design (the table covers the area of 40 to 16,000 watts).

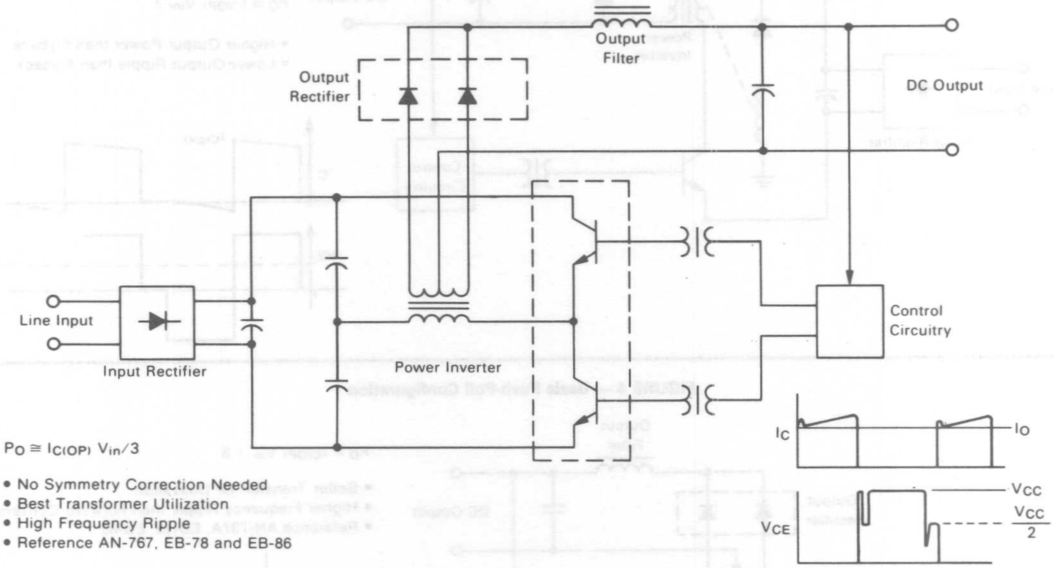
Tables 1 through 3 list devices by $V_{CEO\ (sus)}$ for use in bridge circuits at either 120, 220 or 380 volts. Tables 4 and 5 list devices by V_{CEV} for use in the push-pull, forward and flyback circuits at either 120 or 220 volts. Within each table, the devices are grouped by the output power capability of that circuit, and the equivalent operating current level is also noted. Schematics of these circuits include references to available Motorola literature.

Basic Switching Power Supply Configurations

Minimum device voltage rating recommended for these two circuits:
See Tables 1, 2 and 3, for recommended devices.

V_{in} V_{ac}	V_{DSS} or $V_{CEO(sus)}$ V_{dc}
120	200
220	400
380	600

FIGURE 1 — Basic Half-Bridge Configuration



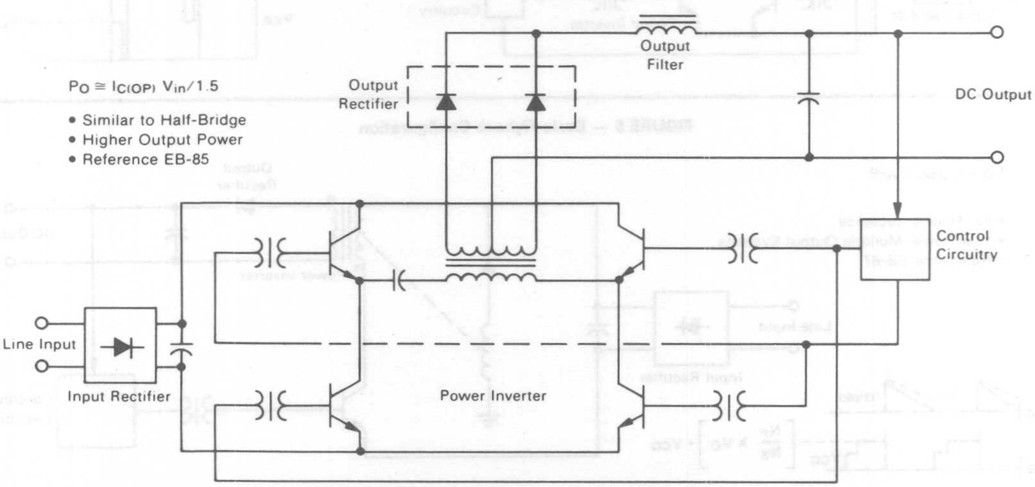
$P_O \cong I_{C(OP)} V_{in}/3$

- No Symmetry Correction Needed
- Best Transformer Utilization
- High Frequency Ripple
- Reference AN-767, EB-78 and EB-86

FIGURE 2 — Basic Full-Bridge Configuration

$P_O \cong I_{C(OP)} V_{in}/1.5$

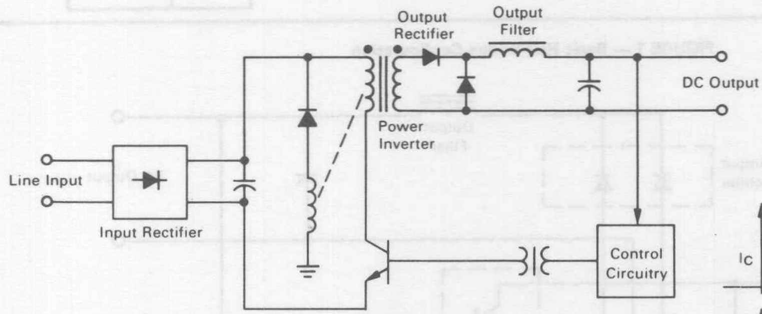
- Similar to Half-Bridge
- Higher Output Power
- Reference EB-85



Minimum recommended device voltage rating for these three circuits:
See Tables 4 and 5, for recommended devices.

V_{in} V_{ac}	V_{DSS} or V_{CEV} V_{dc}
120	450
220	850

FIGURE 3 — Basic Forward Converter



$P_O \cong I_{C(OP)}\ V_{in}/3$

- Higher Output Power than Flyback
- Lower Output Ripple than Flyback

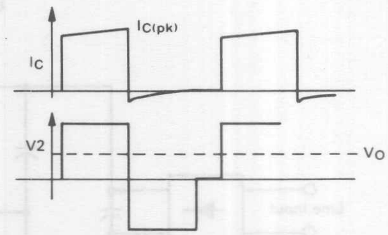
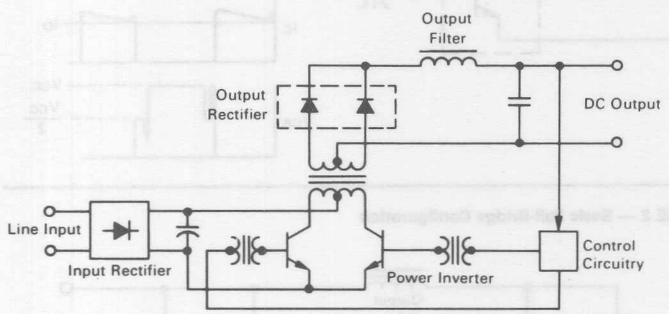


FIGURE 4 — Basic Push-Pull Configuration



$P_O \cong I_{C(OP)}\ V_{in}/1.5$

- Better Transformer Utilization
- Higher Frequency Ripple than Forward Converter
- Reference AN-737A, EB-66, EB-88

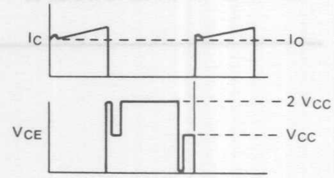
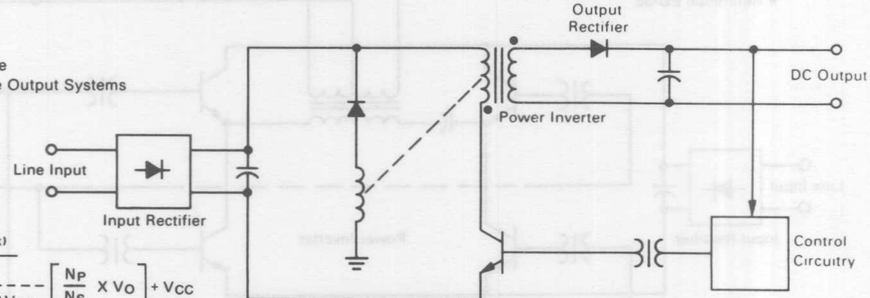


FIGURE 5 — Basic Flyback Configuration

$P_O \cong I_{C(OP)}\ V_{in}/6$

- No Filter Inductance
- Low Power Multiple Output Systems
- Reference EB-87



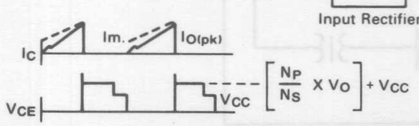


TABLE 1
Circuit: Half and Full Wave Bridge*
Line Voltage: 120 VRMS

OTHER ASSOCIATED REFERENCES:
- AN-778 Mounting Power Devices
- AN-785 Reverse Bias Safe Area
- AN-786 Power Darlington Load Line Considerations
- AN-803 The Effect of Emitter-Base Avalanching on High-Voltage-Power Switching Transistors
- AN-828 The Effects of Base Drive Conditions on RBSPA
- AN-845 New Bipolars Compare Favorably with FETs for Switching Efficiency

Circuit Rating		Bipolar Transistors						TMOS Power MOSFETs			
Output Power*	$I_{D(OP)}$ or $I_{C(OP)}$	Metal—TO-204, TO-66		Plastic TO-220, TO-126, TO-218		Darlington—TO-204		Metal—TO-204		Plastic—TO-220	
(Watts)	(Amps)	Device	V_{CEO}	Device	V_{CEO}	Device	V_{CEO}	Device	V_{DSS}	Device	V_{DSS}
40	1	2N6078 2N3584 2N6234 2N3585 2N6212PNP MJ4360	275 250 275 300 300 300	MJE13002 TIP48 TIP49 TIP50	300 300 350 400			MTM3N35	350	MTP2N20 MTP3N35	200 350
80	2			BUS45P MJE13004	400 300			MTM5N35	350	MTP5N35	350
120	3	MJ6502PNP MJ4380 MJ4400 MJ4381 MJ4401 MJ16004	250 300 300 400 400 450	2N6498 MJE16004 BUS46P	300 450 400			MTP5N20	200	MTP5N20	200
200	5	BUS47 MJ16006 2N6502PNP	400 450 250	MJE13006 MJE5850PNP MJE5851PNP BUS47P**	300 300 350 400	MJ10006	350	MTM12N20	200		
320	8			MJE13008	300						
400	10	BUS48 2N6676 2N6677 2N6678 MJ16010	400 300 350 400 450	BUS48P**	400	MJ10004	350	MTM15N35	350		
800	20	BUS98 MJ16020	400 450			MJ10022 MJ10015	350 400				
1200	30					MJ10020 MJ10021	200 250			MTE60N20***	200
4000	100					MJ10047***	250				
8000	200					MJ10201*	250				

*Case 346-01, JEDEC MO-040 AA **Case TO-218 ***Case 353-01
*NOTE: Power levels shown apply to half-bridge and should be multiplied by 2 if a full bridge is used.

TABLE 2
Circuit: Half and Full Bridge*
Line Voltage: 230 VRMS

Circuit Rating		Bipolar Transistors						TMOS Power MOSFETs			
Output Power	$I_{D(OP)}$ or $I_{C(OP)}$	Metal—TO-204, TO-66		Plastic TO-220, TO-126, TO-218		Darlington—TO-204		Metal—TO-204		Plastic—TO-220	
(Watts)	(Amps)	Device	V_{CEO}	Device	V_{CEO}	Device	V_{CEO}	Device	V_{DSS}	Device	V_{DSS}
80	1	MJ4361	400	MJE13003 TIP50 MJE16001	400 400 450			MTM2N45 MTM2P45	450 450	MTP2N45 MTP2P45	450 450
160	2	MJ4381	400	MJE13005	400			MTM4N45	450	MTP4N45	450
240	3	MJ4401 MJ16004	400 450	BUS46P	400			MTM7N45	450		
400	5	MJ6503PNP MJ16006	400 450	MJE13007 MJE5852PNP BUS47P** BUS47AP** MJH16006**	400 400 400 450 450	MJ1007 BUT50P**	400 500				
640	8			MJE13009	400	MJ10013	550	MTM15N40	400		
800	10	2N6677 BUS48 BUS97 2N6678 BUS48A BUS97A MJ16010	400 400 400 450 450 450 450	BUS48P** BUS48AP**	400 450	MJ10005 MJ10008 BUT13 MJ10009 BUT14 BUT51P** MJ10013 MJ10014 BUT15	400 450 400 500 500 500 550 600 700				
1600	20	BUS98 MJ16016 MJ16020	400 450 450			MJ10023 MJ10015 BUT33 BUT34 MJ10016	400 400 400 500 500				
3500	50					MJ10044***	450				
7000	100					MJ10101*	450				

*NOTE: Power levels shown apply to forward converters and should be multiplied by 2 for push-pull and divided by 2 for flyback.
*Indicates: Case 346-01, JEDEC MO-040AA ***Case 353-01 **Case 218

TABLE 3
Circuit: Half and full Bridge*
Line voltage: 380 VRMS

Circuit Rating		Bipolar Transistors						TMOS Power MOSFETs			
Output Power*	$I_{D(OP)}$ or $I_{C(OP)}$	Metal—TO-204, TO-66		Plastic—TO-220, TO-218		Darlington—TO-204		Metal—TO-204		Plastic—TO-220	
(Watts)	(Amps)	Device	V_{CEO}	Device	V_{CEO}	Device	V_{CEO}	Device	V_{DSS}	Device	V_{DSS}
240	2	MJ8500 MJ12002 MJ8501	700 750 800	MJE8500 MJE12007 MJE8501	700 750 800			MTM3N60	600	MTP3N60	600
360	3	MJ8502 MJ12003 MJ8503	700 750 800	MJE8502 MJE8503	700 800			MTM6N60	600	MTH6N60**	600
480	4	MJ12004	750			MJ10011	700				
600	5	MJ8504 MJ12005 MJ8505	700 750 800								
1200	10					MJ10014 MJ10024 MJ10025 BUT16 BUT36	600 750 850 1000 1000				
3000	25					MJ10041***	850				
6000	50					MJ10051*	850				

*NOTE: Power levels shown apply to forward converters and should be multiplied by 2 for push-pull and divided by 2 for flyback.
*Indicates: Case 346-01, JEDEC MO-040AA ***Case 353-01 **Case 218

TABLE 4
Circuit: Forward, Push-Pull and Flyback*
Line Voltage: 120 VRMS

Circuit Rating		Bipolar Transistors						TMOS Power MOSFETs			
Output Power*	$I_{D(OP)}$ or $I_{C(OP)}$	Metal—TO-204, TO-66		Plastic TO-220, TO-126, TO-218		Darlington—TO-204		Metal—TO-204		Plastic—TO-220-TO218	
(Watts)	(Amps)	Device	V_{CEV}	Device	V_{CEV}	Device	V_{CEV}	Device	V_{DSS}	Device	V_{DSS}
40	1	2N3585	450	MJE13002 MJE13003 TIP50	600 700 400			MTM2N45 MTM2P45	450 450	MTP2N45 MTP2P45	450 450
80	2	MJ4380 MJ4381	600 700	MJE13004 MJE13005	600 700			MTM4N45	450	MTP4N45	450
120	3	2N6306	500	BUS46P	850			MTM7N45	450	MTH7N45** MTH6N55**	450 550
200	5	MJ6503PNP BUS47	450 850	MJE5852PNP MJE5740 MJE13006 MJE5741 MJE13007 MJE5742 BUS47P**	450 600 600 700 700 800 850	MJ10005 MJ10007 BUT50P**	450 500 500				
320	8			MJE13008 MJE13009	600 700			MTM15N45	450		
400	10	2N6676 2N6678 BUS48	500 650 850	BUS48P**	850	MJ10004 MJ10005 BUT13 MJ10008 MJ10013 MJ10014 MJ10009 BUT14 BUT15	450 500 600 650 650 700 750 850 1000				
800	20	BUS98	850			BUT33 MJ10015 MJ10016 BUT34	600 600 700 850				
2000	50					MJ10044***	450				
4000	100					MJ10101*	450				

*NOTE: Power levels shown apply to forward converters and should be multiplied by 2 for push-pull and divided by 2 for flyback.
*Indicates Case 346-01, JEDEC MO-040AA **Indicates Case TO-218 ***Indicates Case 353-1

TABLE 5
Circuit: Forward, Push-Pull and Flyback*
Line Voltage: 220 VRMS

Circuit Rating		Bipolar Transistors						TMOS Power MOSFETs			
Output Power	$I_{D(OP)}$ or $I_{C(OP)}$	Metal—TO-204, TO-66		Plastic—TO-220		Darlington—TO-204		Metal—TO-204		Plastic—TO-220	
(Watts)	(Amps)	Device	V_{CEV}	Device	V_{CEV}	Device	V_{CEV}	Device	V_{DSS}	Device	V_{DSS}
80	1	MJ8500 MJ85001 MJ12002	1200 1400 1500	MJE8500 MJE8501	1200 1400			MTM2N85 MTM2N90	850 900	MTP2N85 MTP2N90	850 900
160	2	MJ8502 MJ8503 MJ12003	1200 1400 1500	MJE8502 MJE8503 MJE12007	1200 1400 1500						
240	3	MJ4401 MJ16002 MJ8504 MJ8505	850 850 1200 1500	BUS46P	850						
400	5	BUS47 MJ16006 BUS47A MJ12004 MJ12005	850 850 1000 1500 1500	BUS47P** BUS47AP**	850 1000	BUT15	1000				
800	10	BUS48 MJ16012 BUS48A	850 850 1000	BUS48P** BUS48AP**	850 1000	MJ10024 MJ10025	1000 1200				

*NOTE: Power levels shown apply to forward converters and should be multiplied by 2 for push-pull and divided by 2 for flyback.
*Indicates Case 346-01, JEDEC MO-040AA   Indicates Case TO-218

TABLE 6
Low Voltage Applications 12-24-48-96 VDC
Line Voltage: 120 VRMS

Circuit Rating	Bipolar Transistors						TMOS Power MOSFETs			
$I_{D(OP)}$ or $I_{C(OP)}$	Metal—TO-204, TO-66		Plastic TO-220, TO-126, TO-218		Darlington—TO-204		Metal—TO-204		Plastic—TO-220	
(Amps)	Device	V_{CEO}	Device	V_{CEO}	Device	V_{CEO}	Device	V_{DSS}	Device	V_{DSS}
1			TIP47 TIP48	250 300					MTP2N18 MTP2N20	180 200
4	BUX42 BUX43	250 325	DH44H-10 D45H-10PNP	80 80			MTM8N18 MTM8N20	180 200	MTP8N18 MTP8N20	180 200
8	BUX13	325	DH44VH-10 DH45VH-10PNP BUV27	100 100 120						
10	BUV11 BUV41 BUV12	200 200 250	BUV26 BUS36 BUS37	90 120 150			MTM20N10	100	MTP20N10	100
15	BUX40 BUV10N	125 125								
20	BUV22	250								
25	BUV21	160								
40	BUS52	250								
50	BUV20 BUS51	125 200							MTE100N06***	60
70	BUS50	125								
100					MJ10047***	250				
200					MJ10201*	250				

*Indicates: Case 346-01, JEDEC MO-040AA **Case TO-218 ***Case 353-01

Bipolar

TABLE 7 — Switchmode Power Transistors

Devices are listed in descending order of $V_{CEO(sus)}$, and I_CCont

V_{CEO} (sus) Volts Min	I_C Cont Amps Max	V_{CEV} Volts Min	Device Type NPN unless otherwise noted	h_{FE} Min/Max	@ I_C Amp	Resistive Switching t_s µs Max	t_f µs Max	@ I_C Amp	f_T MHz Min	Case JEDEC/MOT
1000	24	1400	BUT36##★	5 min	16	6.0	2.5	16		TO-197/1
	12	1400	BUT16##★	5 min	8	3.3	1.5	8		TO-204/1
850	50	900	MJ10050#★	40 min	50	100	35	50		MO-040/346
	50	900	MJ10051##★	40 min	50	10	5	50		MO-040/346
	25	900	MJ10041##★	40 min	25	10	5	25		353/1
	20	1200	MJ10025##★	50/600	20	5	1.8	10		TO-204/1
800	10	1400	MJ8505★	7.5 min	1.5	4	2	5		TO-204/1
	5	1400	MJ8503★	7.5 min	1	4	2	2.5		TO-204/1
	5	1400	MJE8503★	7.5 min	1.0	4	2	2.5		TO-220/221A
	2.5	1400	MJ8501★	7.5 min	0.5	4	2	1		TO-204/1
	2.5	1400	MJE8501★	7.5 min	0.5	4	2	1		TO-220/221A
750	50	900	MJ10052##★	40 min	50	10	5	50		MO-040/346
	20	1000	MJ10024##★	50/600	20	5	1.8	10		TO-204/1
	8	1500	MJ12005	5 min	5		1	5	4 typ	TO-204/1
	5	1500	MJ12004★	2.5 min	4.5		1	4.5	4 typ	TO-204/1
	4	1500	MJ12003	2.5	3		1	3	4 typ	TO-204/1
	2.5	1500	MJ12002★	1.11	2		1	2	4 typ	TO-204/1
		1500	MJE12007★	1.1 min	2		1	2	4 typ	TO-220/221A
700	40	1000	BUT35##★	15 min	24	4.0	1.2	24		TO-197/1
	20	1000	BUT15##★	15 min	12	2.5	0.8	12		TO-204/1
	10	1200	MJ8504★	7.5 min	1.5	4	2	5		TO-204/1
	8	1400	MJ10011#	20 min	4		1	4		TO-204/1
	5	1200	MJ8502★	7.5 min	1	4	2	2.5		TO-204/1
		1200	MJE8502★	7.5 min	1	4	2	2.5		TO-220/221A
	2.5	1200	MJ8500★	7.5 min	0.5	4	2	1		TO-204/1
		1200	MJE8500★	7.5 min	0.5	4	2	1		TO-220/221A
600	15	700	MJ10014##★	10/250	10	2.5	0.8	10		TO-204/1
550	10	650	MJ10013##★	10/250	10	2.5	0.8	10		TO-204/1
500	50	850	BUT34##★	15	32	3.0	1.5	32		TO-197/1
	50	750	MJ10016##★	10 min	40	2.5	1	20		TO-204 Mod/197
	25	850	BUT14##★	15 min	16	2.8	0.8	16		TO-204/1
	20	600	MJ10009##★	30/300	10	2	0.6	10	8**	TO-204/1
	15	850	BUT51P#★	40	5	1.1	0.16	10		TO-218
	8	850	BUT50P#★	30	2	0.75	0.10	5		TO-218
450	100	500	MJ10100#★	60 min	100	25	10	100		MO-040/346
		500	MJ10101##★	60 min	100	3.75	1.25	100		MO-040/346
	50	500	MJ10044##★	60 min	50	3.8	1.5	50		353/1
	30	850	MJ16020	5.0 min	30	2	0.2	20		TO-197/1
		850	MJ16022	7.0 min	30	1.7	0.15	20		TO-197/1
	20	650	MJ10008##★	30/300	10	2	0.6	10	8**	TO-204/1
		850	MJ16014★	5 min	20	2.7	0.35	20		TO-204/197
		850	MJ16016★	7 min	20	2.2	0.25	20		TO-204/197
	15	1000	BUS48A★	8 min	10	2.5	0.35	10		TO-204/1
		1000	BUS48AP★	8 min	10	2.5	0.35	10		TO-218
		850	MJ16010★	5 min	15	1.2 typ	0.2 typ	10		TO-204/1
		850	MJ16012★	7 min	15	0.9 typ	0.15 typ	10		TO-204/1
	8	1000	BUS47A★	7 min	6	2.5	0.3	6		TO-204/1
		1000	BUS47AP★	7 min	6	2.5	0.3	6		TO-218
		850	MJH16008★	7 min	8	2.2	0.25	5		TO-218
		850	MJ16006★	5 min	8	2.5	0.25	5		TO-204/1
		850	MJ16008★	7 min	8	2.2	0.25	5		TO-204/1

★ Designers Data Sheet characterization * t_{off} ** $|h_{fe}|$ @ 1 MHz

Darlington ## Darlington with speed-up diode

(Continued on next page)

Bipolar (continued)

TABLE 7 — Switchmode Power Transistors (continued)

$V_{CEO (sus)}$ Volts Min	I_C Cont Amps Max	V_{CEV} Volts Min	Device Type NPN unless otherwise noted	h_{FE} Min/Max	@ I_C Amp	Resistive Switching t_s µs Max	t_f µs Max	@ I_C Amp	f_T MHz Min	Case JEDEC/MOT
450	5	850	MJH16002★	5 min	5	3	0.3	3		TO-218
		850	MJH16004★	7 min	5	2.7	0.35	3		TO-218
		850	MJ16002★	5 min	5	3	0.3	3		TO-204/1
		850	MJ16004★	8 min	3	2.7	0.35	3		TO-204/1
		850	MJE16002★	5 min	5	3	0.3	3		TO-220/221A
		850	MJE16004★	7 min	5	2.7	0.35	3		TO-220/221A
400	56	600	BUT33##★	20 min	36	3.3	1.6			TO-204 Mod/197
	50	650	MJ10015##★	10 min	40	2.5	1	20		TO-204 Mod/197
	40	600	MJ10023##★	50/600	10	2.5	0.9	20		TO-204Mod/197
	30	850	BUS98★	8 min	20	2.8	0.35	20		TO-204Mod/197
	28	600	BUT13##★	20 min	18	2.6	0.8	18		TO-204/1
	20	500	MJ10001#★	40/400	10	3	1.8	10	10**	TO-204/1
		500	MJ10005#★	40/400	10	1.5	0.5	10	10**	TO-204/1
		450	BUV24★	8 min	12	3.0	0.9	12	8	TO-204/1
	15	850	BUS48★	8 min	10	2.5	0.35	10		TO-204/1
		850	BUS48P★	8 min	10	2.5	0.35	10		TO-218
		650	2N6678	8 min	15	2.5	0.5	15	3	TO-204/1
	12	700	MJE13009★	6/30	8	3	0.7	8	4**	TO-220/221A
	10	500	MJ10003#★	30/300	5	2.5	1	5	10**	TO/204/1
		500	MJ10007##★	30/300	5	1.1	0.5	5	10**	TO-204/1
	9	850	BUS47★	7 min	6	2.5	0.3	6		TO-204/1
		850	BUS47P★	7 min	6	2.5	0.3	6		TO218
		850	MJE5742★	200/400	4	8 typ	2 typ	6		TO-220/221A
		700	MJE13007★	6/30	5	3	0.7	5	4	TO-220/221A
		450	MJ6503-PNP★	15 min	2	2	0.5	4		TO-204/1
		450	MJE5852-PNP★	15 min	2	2	0.5	4		TO-220/221A
	5	850	MJ4401★	7/35	3	4	0.8	3	6	TO-66/80
		850	BUS46P★	7 min	3	2.0	0.3	3		TO-220/221A
	4	700	MJE13005★	6/30	3	3	0.7	3	4	TO-220/221A
		700	MJ4381★	8/40	2	4	0.9	2	4	TO-66/80
	3	850	BUS45P	6 min	2	2.0	0.35	2		TO220/221A
	1.5	700	MJE13003★	5/25	1	4	0.7	1	5	TO-126/77R
		700	MJ4361★	5/25	1	4	0.7	1	4	TO-66/80
	1	500	TIP50★	30/150	0.3				25**	TO-220/221
	0.5	400	MJ4647-PNP	20 min	0.5	0.72*		0.05	40	TO-39/79
350	100	500	MJ10102##★	60 min	100	3.75	1.25	100		MO-040/346
	40	450	MJ10022##★	50/600	10	2.5	0.9	20		TO-204 Mod/197
	20	450	MJ10000#★	40/400	10	3	1.8	10	10**	TO-204/1
		450	MJ10004##★	40/400	10	1.5	0.5	10	10**	TO-204/1
	15	550	2N6677	8 min	15	2.5	0.5	15	3	TO-204/1
	10	450	MJ10002#★	30/300	5	2.5	1	5	10**	TO-204/1
		450	MJ10006##★	30/300	5	1.5	0.5	5	10**	TO-204/1
	8	700	MJE5741#	200/400	4	8 typ	2 typ	6		TO-220/221A
		400	MJE5851-PNP	15 min	2	2	0.5	4		TO-220/221A
	2	400	2N6213PNP	10/100	1	2.5	0.6	1	20	TO-66/80
	1	450	TIP49★	30/150	0.3				25**	TO220/221
325	30	400	BUV23★	8 min	16	1.8	0.4	16	8.0	TO-204 Mod 197
	15	400	BUX13★	8 min	8	2.5	0.8	8	8.0	TO-204/1
	10	400	BUX43	8 min	5	2.2	0.9	5	8.0	TO-204/1
	3	350	2N6235	25/125	1	3.5	0.5	1	20	TO-66/80
300	15	450	2N6676	8 min	15	2.5	0.5	15	3	TO-204/1
	12	600	MJE13008★	6/30	8	3	0.7	8	4**	TO-220/221A
	8	600	MJE13006★	6/30	5	3	0.7	5	4	TO-220/221A
		600	MJE5740	200/400	4	8 typ	2 typ	6		TO-220/221A
		350	MJE5850-PNP★	15 min	2	2	0.5	4		TO-220/221A

★ Designers Data Sheet characterization * t_{off} ** $|h_{fe}|$ @ 1 MHz

Darlington ## Darlington with speed-up diode Note: TO-204 was formerly TO-3

(Continued on next page)

Bipolar (continued)

TABLE 7 — Switchmode Power Transistors (continued)

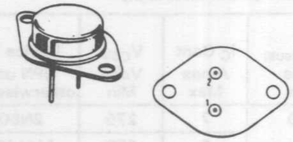
$V_{CEO(sus)}$ Volts Min	I_C Cont Amps Max	V_{CEV} Volts Min	Device Type NPN unless otherwise noted	h_{FE} Min/Max	@ I_C Amp	Resistive Switching t_s µs Max	t_f µs Max	@ I_C Amp	f_T MHz Min	Case JEDEC/MOT
300	7	275	2N6077	12/70	1.2	2.8	0.3	1.2	7	TO-66/80
	5	650	MJ4400★	7/35	3	4	0.8	3	6	TO-66/80
		400	2N6498	10/75	2.5	1.8	0.8	2.5	5	TO-220/221A
	4	600	MJE13004★	6/30	3	3	0.7	3	4	TO-220/221A
		600	MJ4380★	8/40	2	4	0.9	2	4	TO-66/80
	2	500	2N3585	25/100	1	4	3	1	10	TO-66/80
		350	2N6212-PNP	10/100	1	2.5	0.6	1	20	TO-66/80
	1.5	600	MJE13002★	5/25	1	4	0.7	1	5	TO-126/77R
		600	MJ4360★	5/25	1	4	0.7	1	4	TO-66/80
	1	300	2N5345-PNP	25/100	0.5	0.6	0.1	0.5	60	TO-66/80
		400	TIP48★	50/150	0.3				25**	TO-220/221
	0.5	300	MJ4646-PNP	20 min	0.5	0.72*		0.05	40	TO-39/79
275	7	275	2N6078	12/70	1.2	2.8	0.3	1.2	7	TO-66/80
	5	275	2N6234	25/125	1	3.5	0.5	1	20	TO-66/80
250	200	300	MJ10200#★	90 min	200	20	8	200		MO-040/346
		300	MJ10201##★	90 min	200	4	1	200		MO-040/346
	100	300	MJ10047##★	90 min	100	6.5	3	100		353/1
	60	350	MJ10021##★	75 min	15	3.5	0.5	30		TO-204 Mod 197
	40	350	BUS52★	15 min	40	2.0	0.3	40		TO-204 Mod 197
		300	BUV22	10 min	20	1.1	0.35	20	8.0	TO-204 Mod 197
	20	300	BUV12★	10 min	10	1.5	0.5	10		TO-204/1
	15	250	MJ11021#PNP	100 min	15				3#	TO-204/1
		250	MJ11022#	100 min	15				3#	TO-204/1
	12	300	BUX42★	8 min	6	2.0	0.4	6		TO-204/1
	8	400	MJ6502-PNP★	15 min	2	2	0.5	4		TO-204/1
	5	500	MJ3029	30 min	0.4		1	3		TO-204/11
	2	375	2N3584	25/100	1	4	3	1	10	TO-66/80
	1	250	2N5344-PNP	25/100	0.5	0.6	0.1	0.5	60	TO-66580
		250	TIP47★	30/150	0.3				25**	TO-220/221
225	2	275	2N6211	10/100	1	2.5	0.6	1	20	TO-66/80
200	200	300	MJ10202##★	90 min	200	4	1	200		MO-040/346
	60	300	MJ10020#★	75 min	15	3.5	0.5	30		TO-204 Mod 197
	50	300	BUS51★	15 min	50	2.0	0.3	50		TO-204 Mod 197
	40	250	BUV21★	10 min	25	1.8	0.4	25		TO-204/1
	20	250	BUV11★	10 min	12	1.8	0.4	12	8.0	TO-204/1
	15	250	BUX41★	8 min	8	1.5	0.4	8	8.0	TO-204/1
		200	MJ11019#PNP	100 min	15				3#	TO-204/1
		200	MJ11020#	100 min	15				3#	TO-204/1
	2	200	2N5052	25/100	0.75	3.5	1.2	0.75	10	TO-66/80
	0.5	200	MJ4645-PNP	20 min	0.5	0.72*		0.05	40	TO-39/79
150	12	300	BUS37★	30 min	10	0.5	0.06	12	30	TO-220/221
125	70	200	BUS50★	15 min	40	2.0	0.3	40		TO-204 Mod 197
	50	160	BUV20★	10 min	50	1.2	0.25	50		TO-204 Mod 197
	25	160	BUV10N★	10 min	20	1.55	0.45	15		TO-204/1
	20	160	BUX40★	8 min	15	1.0	0.25	15		TO-204/1
120	12	250	BUS36★	30 min	10	0.5	0.06	12	30	TO-220/221
		120	BUV27			1.2	0.4	8		TO-220/221
90	12	90	BUV26			2	0.15	12		TO-220/221
80	15	100	D44VH-10★	20 min	4.0	0.7	0.09	8.0	50	TO-220/221
		100	D45VH-10★PNP	20 min	4.0	0.7	0.09	8.0	50	TO-220/221
	10		D44E-3#★	1000 min	5.0	2.0	0.5	10.0		TO-220/221
			D45E-3#★PNP	1000 min	5.0	2.0	0.5	10.0		TO-220/221
			D44H-10★	20 min	4.0	0.5	0.14	5.0	50	TO-220/221
			D45H-10★PNP	20 min	4.0	0.5	0.14	5.0	40	TO-220/221

★ Designers Data Sheet characterization
Darlington   ## Darlington with speed-up diode   Note: TO-204 was formerly TO-3
* t_{off} ** $|h_{fe}|$ @ 1 MHz

TMOS Power MOSFETs

TABLE 8 — TO-204 (TO-3)
Metal TMOS Power MOSFETs

STYLE 3:
PIN 1. GATE
2. SOURCE
CASE. DRAIN



$V_{(BR)DSS}$ Volts Min	$r_{DS(on)}$ (Ohms) Max	@ I_D Amps	Device	I_D Cont Amps Max	P_D @ T_C = 25°C Watts
1000	10	0.5	MTM1N100	1.0	75
950			MTM1N95		
900	8.0	1.0	MTM2N90	2.0	
850			MTM2N85		
600	2.5	1.5	MTM3N60	3.0	
	1.2	3.0	MTM6N60	6.0	150
550	2.5	1.5	MTM3N55	3.0	75
	1.2	3.0	MTM6N55	6.0	150
500	6.0	1.0	MTM2P50*	2.0	75
	4.0		MTM2N50		
	2.0	2.5	IRF432	4.0	
	1.5		IRF430	4.5	
		2.0	MTM4N50	4.0	
		3.0	2N6762	4.5	
	0.8	3.5	MTM7N50	7.0	150
	0.4	7.5	MTM15N50	15	250
450	6.0	1.0	MTM2P45*	2.0	75
	4.0		MTM2N45		
	2.0	2.5	IRF433, 2N6761	4.0	
	1.5		IRF431	4.5	
		2.0	MTM4N45	4.0	
	0.8	3.5	MTM7N45	7.0	150
	0.40	7.5	MTM15N45	15	250
400	3.3	1.5	MTM3N40	3.0	75
	1.5	3.0	IRF332	4.5	
	1.0		IRF330	5.5	
		2.5	MTM5N40	5.0	
		3.5	2N6760	5.5	
	0.55	4.0	MTM8N40	8.0	150
	0.30	7.5	MTM15N40	15	250
350	3.3	1.5	MTM3N35	3.0	75
	1.5	3.0	IRF333, 2N6759	4.5	
	1.0		IRF331	5.5	
		2.5	MTM5N35	5.0	
	0.55	4.0	MTM8N35	8.0	150
	0.30	7.5	MTM15N35	15	250
250	0.50	5.0	MTM10N25	10	100
200	1.0	2.5	MTM5N20	5.0	75
	0.7	3.5	MTM7N20	7.0	
	0.4	6.0	2N6758	9.0	
	0.4	4.0	MTM8N20	8.0	
	0.35	6.0	MTM12N20	12	100
	0.16	7.5	MTM15N20	15	150
	0.085	10	MTM20N20	20	
180	1.0	2.5	MTM5N18	5.0	75
	0.70	3.5	MTM7N18	7.0	
	0.40	4.0	MTM8N18	8.0	
	0.35	6.0	MTM12N18	12	100
	0.16	7.5	MTM15N18	15	150
	0.085	10	MTM20N18	20	
150	0.70	3.5	MTM7N15	7.0	75
	0.60	5.0	2N6758	9.0	
	0.50	4.0	MTM8N15	8.0	
	0.30	5.0	MTM10N15	10	
	0.25	7.5	MTM15N15	15	150
	0.12	10	MTM20N15	20	
	0.075	12.5	MTM25N15	25	

TABLE 8 — TO-204 (TO-3) (continued)
Metal TMOS Power MOSFETs

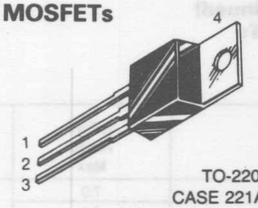
$V_{(BR)DSS}$ Volts Min	$r_{DS(on)}$ @ I_D (Ohms) Max	Amps	Device	I_DCont Amps Max	P_D @ T_C = 25°C Watts
120	0.70	3.5	MTM7N12	7.0	75
	0.50	4.0	MTM8N12	8.0	
	0.30	5.0	MTM10N12	10	
	0.25	7.5	MTM15N12	15	150
	0.12	10	MTM20N12	20	
	0.075	12.5	MTM25N12	25	
100	0.50	4.0	MTM8N10	8.0	75
	0.40		MTM8P10*		
	0.33	5.0	MTM10N10	10	
	0.25	8.0	IRF132	12	
	0.18		IRF130	14	
		6.0	MTM12N10	12	
		9.0	2N6756	14	
	0.15	10	MTM20N10	20	100
	0.07	12.5	MTM25N10	25	150
	0.055	17.5	MTM35N10	35	
80	0.50	4.0	MTM8N08	8.0	75
	0.40		MTM8P08*		
	0.33	5.0	MTM10N08	10	
	0.18	6.0	MTM12N08	12	
	0.15	10	MTM20N08	20	100
	0.07	12.5	MTM25N08	25	150
	0.055	17.5	MTM35N08	35	
60	0.28	5.0	MTM10N06	10	75
	0.25	8.0	IRF133, 2N6755	12	
	0.20	6.0	MTM12N06		
	0.18	8.0	IRF131	14	
	0.16	7.5	MTM15N06	15	
	0.08	12.5	MTM25N06	25	100
	0.055	17.5	MTM35N06	35	150
	0.040	25	MTM50N06	50	
50	0.28	5.0	MTM10N05	10	75
	0.20	6.0	MTM12N06	12	
	0.16	7.5	MTM15N05	15	
	0.08	12.5	MTM25N05	25	100
	0.055	17.5	MTM35N05	35	150
	0.040	25	MTM50N05	50	

*Indicates P-Channel

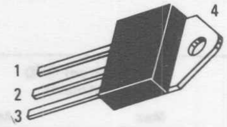
TMOS POWER MOSFETs

TABLE 9 — TO-220 (MTP) — TO-218 (MTH)
Plastic TMOS Power MOSFETs

PIN 1. GATE
2. DRAIN
3. SOURCE
4. DRAIN



TO-220
CASE 221A-02
(MTP and IRF devices in TO-220)



TO-218
CASE 340-01
(MTH devices are in TO-218)

PIN 1. GATE
2. DRAIN
3. SOURCE
4. DRAIN

$V_{(BR)DSS}$ Volts Min	$r_{DS(on)}$ @ I_D (Ohms) Max	Amps	Device	I_DCont Amps Max	P_D @ T_C = 25°C Watts
1000	10	0.5	MTP1N100	1.0	75
950			MTP1N95		
900	8.0	1.0	MTP2N90	2.0	
850			MTP2N85		
600	12	0.5	MTP1N60	1.0	
	2.5	1.5	MTP3N60	3.0	
	1.2	3.0	MTH6N60	6.0	150
550	12	0.5	MTP1N55	1.0	40
	2.5	1.5	MTP3N55	3.0	75
	1.2	3.0	MTH6N55	6.0	150
500	8.0	0.5	MTP1N50	1.0	50
	6.0	2.5	MTP2P50*	2.0	75
	4.0		MTP2N50		
	2.0	1.5	IRF832	4.0	
	1.5		IRF830	4.5	
		2.0	MTP4N50	4.0	
	0.8	3.5	MTH7N50	7.0	150
450	8.0	0.5	MTP1N45	1.0	50
	6.0	1.0	MTP2P45*	2.0	75
	4.0		MTP2N45		
	2.0	2.5	IRF833	4.0	
	1.5		IRF831	4.5	
		2.0	MTP4N45	4.0	
	0.8	3.5	MTH7N45	7.0	150
400	5.0	1.0	MTP2N40	2.0	50
	3.3	1.5	MTP3N40	3.0	75
	1.5	3.0	IRF732	4.5	
	1.0		IRF730	5.5	
		2.5	MTP5N40	5.0	
	0.55	4.0	MTH8N40	8.0	150
350	5.0	1.0	MTP2N35	2.0	50
	3.3	1.5	MTP3N35	3.0	75
	1.5	3.0	IRF733	4.5	
	1.0		IRF731	5.5	
		2.5	MTP4N35	5.0	
	0.55	4.0	MTH8N35	8.0	150
250	0.50	5.0	MTP10N25	10	100
	2.70	1.0	MTP2N25	2.0	50
200	1.8	1.0	MTP2N20	2.0	50
	1.0	2.5	MTP5N20	5.0	75
	0.7	3.5	MTP7N20	7.0	
	0.4	4.0	MTP8N20	8.0	
	0.35	6.0	MTP12N20	12	100
	0.16	7.5	MTH15N20	15	150
180	1.8	1.0	MTP2N18	2.0	50
	1.0	2.5	MTP5N18	5.0	75
	0.7	3.5	MTP7N18	7.0	
	0.4	4.0	MTP8N18	8.0	
	0.35	6.0	MTP12N18	12	100
	0.16	7.5	MTH15N18	15	150
150	1.3	1.5	MTP3N15	3.0	50
	0.7	3.5	MTP7N15	7.0	75
	0.5	4.0	MTP8N15	8.0	
	0.3	5.0	MTP10N15	10	
	0.25	7.5	MTH15N15	15	150
	0.12	10	MTH20N15	20	

*Indicates P-Channel

TABLE 9 — TO-220 (MTP) — TO-218 (MTH)
Plastic TMOS Power MOSFETs (continued)

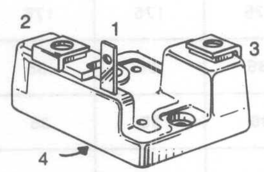
$V_{(BR)DSS}$ Volts Min	$r_{DS(on)}$ @ I_D (Ohms) Max	Amps	Device	I_D Cont Amps Max	P_D @ T_C = 25°C Watts
120	1.3	1.5	MTP3N12	3.0	50
	0.7	3.5	MTP7N12	7.0	75
	0.5	4.0	MTP8N12	8.0	
	0.3	5.0	MTP10N12	10	
	0.25	7.5	MTH15N12	15	150
	0.12	10	MTH20N12	20	
100	0.8	2.0	MTP4N10	4.0	50
	0.5	4.0	MTP8N10	8.0	75
	0.4		MTP8P10*		
	0.33	5.0	MTP10N10	10	
	0.25	8.0	IRF532	12	
	0.18		IRF530	14	
		6.0	MTP12N10	12	
	0.15	10	MTP20N10	20	100
	0.07	12.5	MTH25N10	25	150
80	0.8	2.0	MTP4N08	4.0	50
	0.5	4.0	MTP8N08	8.0	75
	0.4		MTP8P08*		
	0.33	5.0	MTP10N08	10	
	0.18	6.0	MTP12N08	12	
	0.15	10	MTP20N08	20	100
	0.07	12.5	MTH25N08	25	150
60	0.8	2.5	MTP5N06	5.0	50
	0.28	5.0	MTP10N06	10	75
	0.25	8.0	IRF533	12	
	0.20	6.0	MTP12N06		
	0.18	8.0	IRF531	14	
	0.16	7.5	MTP15N06	15	
	0.08	12.5	MTP25N06	25	100
	0.055	17.5	MTH35N06	35	150
50	0.8	2.5	MTP5N05	5.0	50
	0.28	5.0	MTP10N05	10	75
	0.20	6.0	MTP12N05	12	
	0.16	7.5	MTP15N05	15	
	0.10	6.0	BUZ10	12	
	0.08	12.5	MTP25N05	25	100
	0.055	17.5	MTH35N05	35	150

*Indicates P-Channel

TABLE 10 — CASE 353
TMOS POWER MOSFETs ENERGY MANAGEMENT SERIES

CASE 353-01

PIN 1. GATE
2. SOURCE
3. DRAIN
4. DRAIN



V_{DSS} (Volts)	$r_{DS(on)}$ @ I_D (Ohms)	(Amp)	Device	$I_{D(Cont)}$ (Amps)	P_D @ T_C = 25°C (Watts)
200	0.048	30	MTE60N20	60	250
180			MTE60N18		
150	0.038	32.5	MTE65N15	65	
120			MTE65N12		
100	0.028	37.5	MTE75N10	75	
80			MTE75N08		
60	0.018	50	MTE100N06	100	
50			MTE100N05		

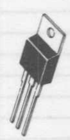
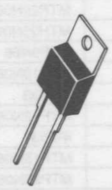
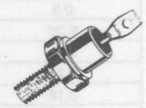
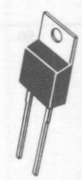
Rectifiers

- **Ultrafast Recovery**
- **Schottky**
- **Fast Recovery**
- **Applications**

Switchmode Rectifiers by Motorola have been developed to provide the designer a broad combination of current, voltage, speed, package and cost. Ultrafast, Schottky and Fast Recovery are all available in different package styles and materials to meet every design requirement.

Ultrafast Recovery Rectifiers

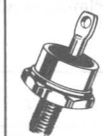
Ultrafast Rectifiers utilize many of the state-of-the-art techniques used to manufacture Schottky rectifiers and IC's to provide a more efficient unit than a conventional 200 nanosecond fast recovery rectifier. Low V_F and I_R plus typical switching speeds of 25-35 nanoseconds make these units ideal for use in switching power supplies.

I_O	I_O AVERAGE FORWARD RECTIFIED CURRENT (A)							
	1	6	7		8	12	15	15
CASE	59-03	221A-02	221B-01			56-02		221B-01
								
V_{RRM}	DO41	TO220AB*	TO220AC			DO4		TO220AC
50	MUR105	MUR605CT	BYW29-50	BYW80-50	MUR805	BYW30-50	BYW81-50	MUR1505
100	MUR110	MUR610CT	BYW29-100	BYW80-100	MUR810	BYW30-100	BYW81-100	MUR1510
150	MUR115	MUR615CT	BYW29-150	BYW80-150	MUR815	BYW30-150	BYW81-150	MUR1515
200	MUR120	MUR620CT	BYW29-200	BYW80-200	MUR820	BYW30-200	BYW81-200	
I_{FSM} (A)	35	75	100	100	100	200	200	150
T_j or T_c at rated I_o (°C)	135	150	150	150	150	120	120	150
T_j max.	175	175	175	175	175	150	150	175
V_F max. at I_{FAV} (V)	.85	.85	.85	.85	.85	.85	.85	.85
T_{rr} max. (ns)	35	35	35	35	35	35	35	35
Thread	—	—	—	—	—	UNF	UNF	—

* I_o is total device output

NOTE: Contact sales office for 400 V and 600 V Ultrafast rectifiers.

Ultrafast Rectifiers (continued)

I_O	I_O AVERAGE FORWARD RECTIFIED CURRENT (A)						
	16			25		30	35
CASE	221A-02			56-02		340-01	257-01
							
V_{RRM}	TO220AB*			DO4		TO-218AC*	DO5
50	BYW51-50	MUR1605CT	BYV32-50	BYW31-50	BYW77-50	MUR3005PT	BYW92-50
100	BYW51-100	MUR1610CT	BYV32-100	BYW31-100	BYW77-100	MUR3010PT	BYW92-100
150	BYW51-150	MUR1615CT	BYV32-150	BYW31-150	BYW77-150	MUR3015PT	BYW92-150
200	BYW51-200	MUR1620CT	BYW32-200	BYW31-200	BYW77-200	MUR3020PT	BYW92-200

I_{FSM} (A)	100	100	150	320	500	200	500
T_j or T_c at rated I_o (°C)	130	150	120	120	115		150
T_j max.	175	175	150	150	150	150	150
V_F max. at I_{FAV} (V)	.85	.85	.85	.85	.85	.85	.92
T_{rr} max. (ns)	35	35	35	50	50	35	50
Thread	—	—	—	UNF	UNF		UNF

* I_o is total device output

NOTE: Contact sales office for 400 V and 600 V Ultrafast rectifiers.

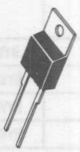
Schottky Rectifiers

Schottky Rectifiers are offered with two different barrier metals:

Chrome provides the lowest forward voltage drop but is limited to an operating temperature of 125°C and has a high leakage current.

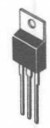
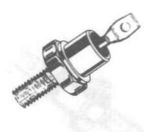
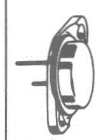
Platinum allows operation at junction temperatures of 150°C and 175°C with a leakage current several orders of magnitude lower than chrome. Here the trade-off is a higher forward voltage.

All Motorola Schottky rectifier chips have a guardring that virtually eliminates dv/dt and transient voltage problems.

I_O	I_O AVERAGE FORWARD RECTIFIED CURRENT (A)									
	1		3			5	8	7	10	15
CASE	59-04		267		60			221B-01		221A-02
										
V_{RRM}							**	TO220AC		TO220AB*
20	1N5817	MBR120P	1N5820	MBR320P	MBR320M	1N5823	BYS08-20			
30	1N5818	MBR130P	1N5821	MBR330P	MBR330M	1N5824	BYS08-30			
35								MBR735	MBR1035	MBR1535CT
40	1N5819	MBR140P	1N5822	MBR340P	MBR340M	1N5825				
45							BYS08-45	MBR745	MBR1045	MBR1545CT
50							BYS08-50			
I_{FSM} (A)	25	25	80	80	500	500	400	150	150	150
T_j or T_c at rated I_o (°C)	90	80	95	85	90	80	100	105	135	105
T_j max.	125	125	125	125	125	125	150	150	150	150
Barrier Metal	CHROME						PLATINUM			
V_F max. at I_o (V)	.60	.65	.53	.60	.45	.38	.47	.57	.57	.72

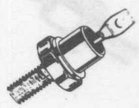
* I_o is total device output ** N° CECC 86-819-029

Schottky Rectifiers (continued)

I_O	I_O AVERAGE FORWARD RECTIFIED CURRENT (A)									
	15		16	20	25			30		
CASE	56-02		221B-01	221A-02	56-02			221A-02	11-03	340-01
										
V_{RRM}	DO4		TO220AC	TO220AB*	DO4			TO220AB	TO3*	TO-218AC*
20	MBR1520	1N5826			MBR2520	1N5829				
30	MBR1530	1N5827			MBR2530	1N5830	1N6095			
35	MBR1535		MBR1635	MBR2035CT	MBR2535			MBR2535CT	MBR3035CT	MBR3035PT
40	MBR1540	1N5828			MBR2540	1N5831	1N6096		SD241	
45			MBR1645	MBR2045CT				MBR2545CT	MBR3045CT	MBR3045PT
I_{FSM} (A)	500	500	300	150	500	800	400	300	400	200
T_j or T_c at rated I_o (°C)	80	80	125	135	80	85	70	125	105	
T_j max.	125	125	150	150	125	125	125	150	150	150
Barrier Metal	CHROME		PLATINUM		CHROME		PLATINUM			
V_F max. at I_o (V)	.55	.50	.60	.72	.55	.48	.58	.73	.65	.72

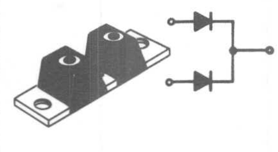
* I_o is total device output

Schottky Rectifiers (continued)

I_O	I_O AVERAGE FORWARD RECTIFIED CURRENT (A)						
	30	35		40		50	60
CASE	56-02			257-01			43-02
							
V_{RRM}	DO4			DO5			DO21
20		MBR3520	BYS35-20**	1N5832	MBR4020		
30		MBR3530	BYS35-30**	1N5833	MBR4030	1N6097	
35		MBR3535			MBR4035		MBR6035PF
40	SD41			1N5834	MBR4040	1N6098	
45		MBR3545	BYS35-45**				MBR6045PF
50			BYS35-50*				
I_{FSM} (A)	600	600	600	800	800	800	800
T_j or T_c at rated I_o (°C)	105	90	100	75	70	70	100
T_j max.	150	150	150	125	125	150	150
Barrier Metal	PLATINUM			CHROME		PLATINUM	
V_F max. at I_o (V)	.55	.55	.60	.59	.63	.60	.65

* I_o is total device output   ** N° CECC 86-819-029

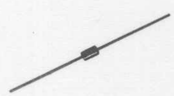
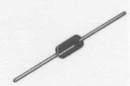
Schottky Rectifiers (continued)

I_O	I_O AVERAGE FORWARD RECTIFIED CURRENT (A)								
	60		65	75		80	120	200	300
CASE	257-01  DO5						357B-01  Power Tap ®		
V_{RRM}									
30		BYS60-30**		BYS75-30***					
35	MBR6035		MBR6535		MBR7535	MBR8035	MBR12035CT*	MBR20035CT*	MBR30035CT
40	SD51								
45	MBR6045	BYS60-45**	MBR6545	BYS75-45***	MBR7545	MBR8045	MBR12045CT*	MBR20045CT*	MBR30045CT
50		BYS60-50**		BYS75-50***					
I_{FSM} (A)	800	800	800	1000	1000	1000	800	1500	2500
T_j or T_C at rated I_o (°C)	100	100	120	100	90	120	120	120	140
T_j max.	150	150	175	150	150	175	175	175	175
Barrier Metal	PLATINUM						PLATINUM		
V_F at rated I_o (V)	.62	.68	.62	.64	.60	.59	.60	.80	0.8

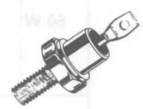
* I_o is total device output ** N° CECC 86-819-031 *** N° CECC 86-819-032

Fast Recovery Rectifiers

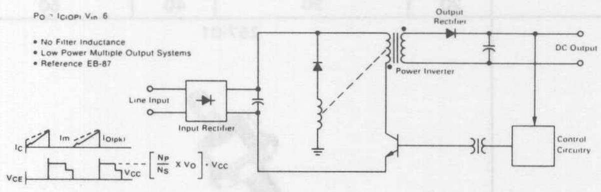
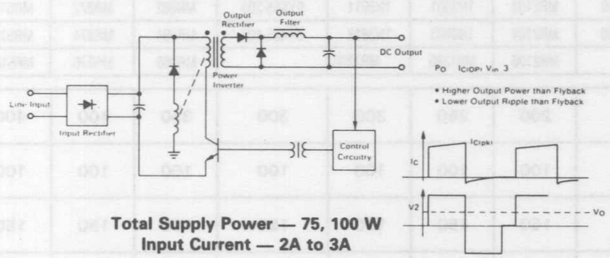
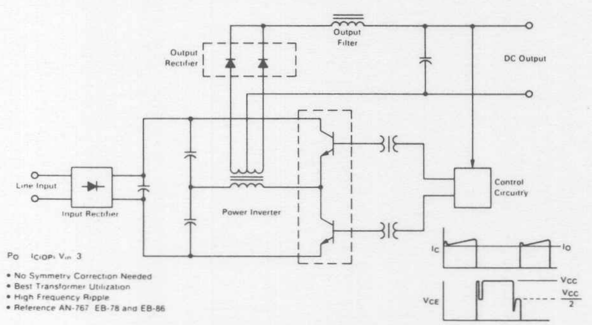
Fast Recovery Rectifiers offer cost effective alternatives with a more mature process technique.

I_O	I_O AVERAGE FORWARD RECTIFIED CURRENT (A)										
	1					3					5
CASE	59-04					267				60	194
V_{RRM}											
50	MR810	1N4933			TSA4933	MR910	MR850		TSA850	MR830	MR820
100	MR811	1N4934	BY196		TSA4934	MR911	MR851	BY296	TSA851	MR831	MR821
200	MR812	1N4935	BY197		TSA4935	MR912	MR852	BY297	TSA852	MR832	MR822
400	MR814	1N4936	BY198	BYX55-350	TSA4936	MR914	MR854	BY298	TSA854	MR834	MR824
600	MR816	1N4937		BYX55-600	TSA4937	MR916	MR856		TSA856	MR836	
800	MR817					MR917					
1000	MR818					MR918					
I_{FSM} (A)	30	30	40	40	30	100	100	100	100	100	350
T_j max. (°C)	150	150	150	125	150	175	175	150	175	150	175
T_{rr} (ns) 1A/30V 50A/us	750	200	200	200	100	750	200	200	100	200	200

Fast Recovery Rectifiers (continued)

I_O	I_O AVERAGE FORWARD RECTIFIED CURRENT (A)									
	6	12			20	30		40	50	
CASE	56-02				257-01					
										
V_{RRM}	DO4				DO5					
50	1N3879	1N3889	BYX61-50		1N3899	1N3909	BYX65-50	MR860	MR870	
100	1N3880	1N3890	BYX61-100		1N3900	1N3910	BYX65-100	MR861	MR871	
200	1N3881	1N3891	BYX61-200	MR2102	1N3901	1N3911	BYX65-200	MR862	MR872	MR5102
400	1N3883	1N3893	BYX61-400	MR2104	1N3903	1N3913	BYX65-400	MR864	MR874	MR5104
600	MR1366	MR1376		MR2106	MR1386	MR1396		MR866	MR876	MR5106
I_{FSM} (A)	150	200	150	200	250	300	300	350	400	400
T_j or T_c at rated I_o (°C)	100	100	100	100	100	100	100	100	100	100
T_J max.	150	150	150	150	150	150	150	150	150	150
T_{RR} max. (ns)	200	200	100	100	200	200	200	200	200	100

Rectifier Application

Typical Circuit	Total Supply Power	Input Rectifiers — Standard Recovery for Line Voltage Operation: Input Current	Suggested Devices*: Type	I_O	V_R*
BASIC FLYBACK CONFIGURATION  **Total Supply Power — 10, 50, 75 W** **Input Current — < 1A to 2A**	10 W	<1 A	1N4004 MDA920A6	1 A 1.5 A	400 V 400 V
	50 W	1A	1N4004 MDA920A6	1 A 1.5 A	400 V 400 V
	75 W	2 A	MR504 1N5404 MDA204	3 A 3 A 2 A	400 V 400 V 400 V
BASIC FORWARD CONVERTER  **Total Supply Power — 75, 100 W** **Input Current — 2A to 3A**	75 W	2 A	MR504 1N5404 MDA204	3 A 3 A 2 A	400 V 400 V 400 V
	100 W	3 A	MR504 1N5404 MDA970A5	3 A 3 A 4 A	400 V 400 V 400 V
BASIC HALF-BRIDGE CONFIGURATION  **Total Power Supply — 250, 100, 2500 W** **Input Current — 6A to 25A**	250 W	6 A	MR754 1N1204, A, B, C MR1124	6 A 12 A 12 A	400 V 400 V 400 V
	1000 W	12 A	1N1204, A, B, C MR1124 BYW24	12 A 12 A .15A	400 V 400 V 400V
	2500 W	25 A	BYW25-400 BYW64 1N1183, A	25 A 35 A 40 A	400 V 400 V 400 V

Rectifier Application (continued)

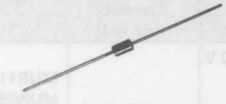
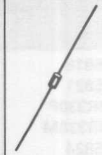
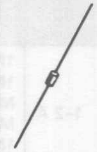
Output Rectifiers											
Schottky For 2.0 V and 5.0 V Outputs				Fast Recovery (200 ns) For ≥5.0 V Outputs				Ultrafast Recovery (35 ns) for ≥5.0 V Outputs			
Output Current	Suggested Devices*			Output Current	Suggested Devices*			Output Current	Suggested Devices*		
	Type	I_O	V_R*		Type	I_O	V_R*		Type	I_O	V_R*
1–2 A	1N5818 1N5821 MBR330P MBR330M 1N5824 MBR735	1 A 3 A 3 A 3 A 5 A 7.5 A	30 V 30 V 30 V 30 V 30 V 35 V	<1/2 A	1N4934	1 A	100 V	<1/2 A	MUR110 MUR860	1 A 8A	100 V 600 V
5–10 A	1N5827 MBR1530 1N5830 1N6095 MBR735 MBR1035 MBR1635	15 A 15 A 25 A 25 A 7.5 A 10 A 16 A	30 V 30 V 30 V 30 V 35 V 35 V 35 V	1/2–1-1/2 A	1N4934 MR851 MR831 MR801	1 A 1 A 3 A 3 A	100 V 100 V 100 V 100 V	1/2–1-1/2 A	MUR110 MUR810 MUR860	1 A 2.5 A @ T_A 8 A	100 V 100 V 600 V
10–15 A	1N5830 MBR2535 SD41 MBR3535 MBR1635 MBR1035	25 A 25 A 30 A 35 A 15 A 10 A	30 V 35 V 35 V 35 V 35 V 35 V	1-1/2–2-1/2 A	MR851 MR821 MR831 MR801	3 A 5 A 3 A 3 A	100 V 100 V 100 V 100 V	1-1/2 A–2-1/2 A	MUR810, R MUR810, R MUR860	2.5 A @ T_A 8 A @ T_C 8 A	100 V 100 V 600 V
8–16 A	1N5827 MBR1530 1N5830 1N6095 MBR1535CT MBR2035CT MBR2535CT	15 A 15 A 25 A 25 A 15 A 20 A 30 A	30 V 30 V 30 V 30 V 35 V 35 V 35 V	2–2-1/2 A	 MR851 MR801	 3 A 3 A	100 V 100 V	2–2-1/2 A	MUR810, R MUR1610CT, R	8 A @ T_C 16 A @ T_C	100 V 100 V
10–20 A	1N5827 MBR1530 1N5830 1N6095 MBR1535CT MBR2535CT	15 A 15 A 25 A 25 A 15 A 30 A	30 V 30 V 30 V 30 V 35 V 35 V	2–2-1/2 A	MR851 MR801	3 A 3 A	100 V 100 V	2–2-1/2 A	MUR810, R MUR1610CT, R	8 A @ T_C 16 A @ T_C	100 V 100 V
30–50 A	1N5830 SD41 1N6095 MBR3535 MBR12035CT	25 A 30 A 25 A 35 A 100 A	30 V 35 V 30 V 35 V 35 V	2–8 A	MR851 MR821 1N3880, A	3 A 5 A 6 A	100 V 100 V 100 V	2–8 A	MUR810, R MUR1610CT, R MUR1510, R	8 A 16 A 15 A	100 V 100 V 100 V
200 A	SD51 MBR6035 MBR7535 1N6097 (In Parallel) MBR20035CT MBR30035CT	60 A 60 A 75 A 50 A 200 A 300 A	35 V 35 V 35 V 30 V 35 V 35 V	40 A	1N3900 1N3910	20 A 30 A	100 V 100 V	40 A	CONSULT FACTORY		
550 A	SD51 MBR6035 MBR7535 1N6097 (In Parallel) MBR20035CT MBR30035CT	60 A 60 A 75 A 50 A 200 A 300 A	35 V 35 V 35 V 30 V 35 V 35 V	100 A	MR871	50 A	100V	100 A	CONSULT FACTORY		

*These suggested devices are the most common used; see the full Selector Guide for other voltages.

Zener Diodes

Zener and Avalanche Regulator Diodes for Switchmode Control & Protection

For detailed parameter and tolerance information, see appropriate Motorola data sheet.

Nominal Zener Voltage	250 mW Low Level Cathode = Polarity Mark	250 mW Low Noise Cathode = Polarity Mark	400 mW Low Noise Low Leakage Cathode = Polarity Mark	500 mW Cathode = Polarity Mark			1 WATT Cathode = Polarity Mark	1.5 WATT Cathode = Polarity Mark	5 WATT Cathode = Polarity Mark
	Case 299-02				Glass DO-204AH (DO-35)		 Glass Case 59 (DO-41)	 Surmetic 30 Case 59 (DO-41)	 Surmetic 40 Case 17
1.8	1N4678	1N4614							
2.0	1N4679	1N4615							
2.2	1N4680	1N4616							
2.4	1N4681	1N4617		1N4370	1N5221	1N5985A			
2.7	1N4682	1N4618		1N4371	1N5223	1N5986A			
3.0	1N4683	1N4619		1N4372	1N5225	1N5987A			
3.3	1N4684	1N4620	1N5518A	1N746	1N5226	1N5988A	1N4728	1N5913A	1N5333A
3.6	1N4685	1N4621	1N5519A	1N747	1N5227	1N5989A	1N4729	1N5914A	1N5334A
3.9	1N4686	1N4622	1N5520A	1N748	1N5228	1N5990A	1N4730	1N5915A	1N5335A
4.3	1N4687	1N4623	1N5521A	1N749	1N5229	1N5991A	1N4731	1N5916A	1N5336A
4.7	1N4688	1N4624	1N5522A	1N750	1N5230	1N5992A	1N4732	1N5917A	1N5337A
5.1	1N4689	1N4625	1N5523A	1N751	1N5231	1N5993A	1N4733	1N5918A	1N5338A
5.6	1N4690	1N4626	1N5524A	1N752	1N5232	1N5994A	1N4734	1N5919A	1N5339A
6.2	1N4691	1N4627	1N5525A	1N753	1N5234	1N5995A	1N4735	1N5920A	1N5341A
6.8	1N4692	1N4099	1N5526A	1N754 1N957A	1N5235	1N5996A	1N4736	1N5921A	1N5342A
7.5	1N4693	1N4100	1N5527A	1N755 1N958A	1N5236	1N5997A	1N4737	1N5922A	1N5343A
8.2	1N4694	1N4101	1N5528A	1N756 1N959A	1N5237	1N5998A	1N4738	1N5923A	1N5344A
8.7	1N4695	1N4102			1N5238				1N5345A
9.1	1N4696	1N4103	1N5529A	1N757 1N960A	1N5239	1N5999A	1N4739	1N5924A	1N5346A
10	1N4697	1N4104	1N5530A	1N758 1N961A	1N5240	1N6000A	1N4740	1N5925A	1N5347A
11	1N4698	1N4105	1N5531A	1N962A	1N5241	1N6001A	1N4741	1N5926A	1N5348A
12	1N4699	1N4106	1N5532A	1N759 1N963A	1N5242	1N6002A	1N4742	1N5927A	1N5349A
13	1N4700	1N4107	1N5533A	1N964A	1N5243	1N6003A	1N4743	1N5928A	1N5350A
14	1N4701	1N4108	1N5534A		1N5244				1N5351A
15	1N4702	1N4109	1N5535A	1N965A	1N5245	1N6004A	1N4744	1N5929A	1N5352A
16	1N4703	1N4110	1N5536A	1N966A	1N5246	1N6005A	1N4745	1N5930A	1N5353A
17	1N4704	1N4111	1N5537A		1N5247				1N5354A
18	1N4705	1N4112	1N5538A	1N967A	1N5248	1N6006A	1N4746	1N5931A	1N5355A
19	1N4706	1N4113	1N5539A		1N5249				1N5356A
20	1N4707	1N4114	1N5540A	1N968A	1N5250	1N6007A	1N4747	1N5932A	1N5357A
22	1N4708	1N4115	1N5541A	1N969A	1N5251	1N6008A	1N4748	1N5933A	1N5358A
24	1N4709	1N4116	1N5542A	1N970A	1N5252	1N6009A	1N4749	1N5934A	1N5359A
25	1N4710	1N4117	1N5543A		1N5253				1N5360A
27	1N4711	1N4118		1N971A	1N5254	1N6010A	1N4750	1N5935A	1N5361A
28	1N4712	1N4119	1N5544A		1N5255				1N5362A
30	1N4713	1N4120	1N5545A	1N972A	1N5256	1N6011A	1N4751	1N5936A	1N5363A
33	1N4714	1N4121	1N5546A	1N973A	1N5257	1N6012A	1N4752	1N5937A	1N5364A
36	1N4715	1N4122		1N974A	1N5258	1N6013A	1N4753	1N5938A	1N5365A
39	1N4716	1N4123		1N975A	1N5259	1N6014A	1N4754	1N5939A	1N5366A
43	1N4717	1N4124		1N976A	1N5260	1N6015A	1N4755	1N5940A	1N5367A
47		1N4125		1N977A	1N5261	1N6016A	1N4756	1N5941A	1N5368A
51		1N4126		1N978A	1N5262	1N6017A	1N4757	1N5942A	1N5369A
56		1N4127		1N979A	1N5263	1N6018A	1N4758	1N5943A	1N5370A
60		1N4128			1N5264				1N5371A
62		1N4129		1N980A	1N5265	1N6019A	1N4759	1N5944A	1N5372A
68		1N4130		1N981A	1N5266	1N6020A	1N4760	1N5945A	1N5373A
75		1N4131		1N982A	1N5267	1N6021A	1N4761	1N5946A	1N5374A
82		1N4132		1N983A	1N5268	1N6022A	1N4762	1N5947A	1N5375A
87		1N4133			1N5269				1N5376A
91		1N4134		1N984A	1N5270	1N6023A	1N4763	1N5948A	1N5377A
100		1N4135		1N985A	1N5271	1N6024A	1N4764	1N5949A	1N5378A
110				1N986A	1N5272	1N6025A	♦1M110ZS10	1N5950A	1N5379A
120				†1N987A	1N5273#		♦1M120ZS10	1N5951A	1N5380A
130				†1N988A	1N5274#		♦1M130ZS10	1N5952A	1N5381A
140					1N5275#				
150				†1N989A	1N5276#		♦1M150ZS10	1N5953A	1N5383A
160				†1N990A	1N5277#		♦1M160ZS10	1N5954A	1N5384A
170					1N5278#		♦1M170ZS10		1N5385A
180				†1N991A	1N5279#		♦1M180ZS10	1N5955A	1N5386A
200				†1N992A	1N5281#		♦1M200ZS10	1N5956A	1N5388A

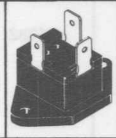
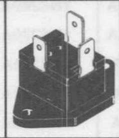
▱ JAN/JANTX(V) available, ±5% only.
†1N987-1N992 supplied in DO-7 glass package.
#1N5273-1N5281 supplied in Surmetic DO-7 plastic package.
♦1M110ZS10 Series supplied in Surmetic (Plastic) DO-41 package.

Mosorb Transient Suppressors

System DC Voltage or System Peak Voltage	600 Watts Peak Pulse Power @ 1.0 ms Surmetic 40 Case 17		1500 Watts Peak Pulse Power @ 1.0 ms Case 41					
	Unidirectional	Bidirectional	Unidirectional	Bidirectional	Unidirectional	Bidirectional	Unidirectional	Bidirectional
5.0			1N5908		1N6373		ICTE-5	
6.0	P6KE6.8A		1N6267A					
6.5	P6KE7.5A	P6KE7.5CA	1N6268A	1.5KE7.5CA				
7.0	P6KE8.2A	P6KE8.2CA	1N6269A	1.5KE8.2CA				
8.0	P6KE9.1A	P6KE9.1CA	1N6270A	1.5KE9.1CA	1N6374	1N6382	ICTE-8	ICTE-8C
8.5	P6KE10A	P6KE10CA	1N6271A	1.5KE10CA				
9.0	P6KE11A	P6KE11CA	1N6272A	1.5KE11CA				
10	P6KE12A	P6KE12CA	1N6273A	1.5KE12CA	1N6375	1N6383	ICTE-10	ICTE-10C
11	P6KE13A	P6KE13CA	1N6274A	1.5KE13CA				
12					1N6376	1N6384	ICTE-12	ICTE-12C
13	P6KE15A	P6KE15CA	1N6275A	1.5KE15CA				
14	P6KE16A	P6KE16CA	1N6276A	1.5KE16CA				
15					1N6377	1N6385	ICTE-15	ICTE-15C
16	P6KE18A	P6KE18CA	1N6277A	1.5KE18CA				
17	P6KE20A	P6KE20CA	1N6278A	1.5KE20CA				
18	P6KE22A	P6KE22CA	1N6279A	1.5KE22CA	1N6378	1N6386	ICTE-18	ICTE-18C
20	P6KE24A	P6KE24CA	1N6280A	1.5KE24CA				
22					1N6379	1N6387	ICTE-22	ICTE-22C
24	P6KE27A	P6KE27CA	1N6281A	1.5KE27CA				
26	P6KE30A	P6KE30CA	1N6282A	1.5KE30CA				
28	P6KE33A	P6KE33CA	1N6283A	1.5KE33CA				
30	P6KE36A	P6KE36CA	1N6284A	1.5KE36CA				
33	P6KE39A	P6KE39CA	1N6285A	1.5KE39CA				
36	P6KE43A	P6KE43CA	1N6286A	1.5KE43CA	1N6380	1N6388	ICTE-36	ICTE-36C
40	P6KE47A	P6KE47CA	1N6287A	1.5KE47CA				
43	P6KE51A	P6KE51CA	1N6288A	1.5KE51CA				
45					1N6381	1N6389	ICTE-45	ICTE-45C
48	P6KE56A	P6KE56CA	1N6289A	1.5KE56CA				
54	P6KE62A	P6KE62CA	1N6290A	1.5KE62CA				
58	P6KE68A	P6KE68CA	1N6291A	1.5KE68CA				
64	P6KE75A	P6KE75CA	1N6292A	1.5KE75CA				
70	P6KE82A	P6KE82CA	1N6293A	1.5KE82CA				
78	P6KE91A	P6KE91CA	1N6294A	1.5KE91CA				
85	P6KE100A	P6KE100CA	1N6295A	1.5KE100CA				
90	P6KE110A	P6KE110CA	1N6296A	1.5KE110CA				
100	P6KE120A	P6KE120CA	1N6297A	1.5KE120CA				
110	P6KE130A	P6KE130CA	1N6298A	1.5KE130CA				
120	P6KE150A	P6KE150CA	1N6299A	1.5KE150CA				
130	P6KE160A	P6KE160CA	1N6300A	1.5KE160CA				
140	P6KE170A	P6KE170CA	1N6301A	1.5KE170CA				
150	P6KE180A	P6KE180CA	1N6302A	1.5KE180CA				
165								
170	P6KE200A	P6KE200CA	1N6303A	1.5KE200CA				

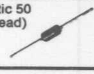
Thyristors for Switchmode Applications

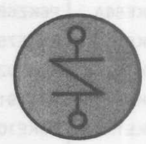
Crowbar SCRs

		PEAK CAPACITOR DISCHARGE CURRENT (1)						
		300 AMPS			750 AMPS		850 AMPS	1700 AMPS
								
		Case 86 Style 1	Case 342-01 Style 1	Case 221A-02 Style 3		Case 342-01 Style 1	Case 175-03 Style 1	Case 263-04 Style 1
V_{DRM} or V_{RRM}	25 V	MCR67-1	MCR568-1	MCR68-1	MCR69-1	MCR569-1	MCR70-1	MCR71-1
	50 V	MCR67-2	MCR568-2	MCR68-2	MCR69-2	MCR569-2	MCR70-2	MCR71-2
	100 V	MCR67-3	MCR568-3	MCR68-3	MCR69-3	MCR569-3	MCR70-3	MCR71-3
	400 V	MCR67-6	MCR568-6	MCR68-6	MCR69-6	MCR569-6	MCR70-6	MCR71-6
ELECTRICAL CHARACTERISTICS Maximum or Min/Max	$I_{T(RMS)}$ (AMPS)	12	12	12	25	25	35	55
	I_{GT} (mA) @ 25°C Min/Max	2/30	2/30	2/30	2/30	2/30	2/30	2/30
	V_{GT} (V) @ 25°C	1.5	1.5	1.5	1.5	1.5	1.5	1.5
	I_H (mA) @ 25°C Min/Max	3/50	3/50	3/50	3/50	3/50	3/50	3/50
	I_L (mA) @ 25°C	60	60	60	60	60	60	60

(1) The peak capacitor discharge current is for t_w = 1.0 ms. t_w is defined as 5 time constants of an exponentially decaying current pulse.

Source: Switchmode Selector Guide 82

SIDAC					
Package	Device Type	Vs Volts Min	Vs Volts Max	IRMS	ISURGE
Surmetic 50 (Axial lead) 	MK1V-125 MK1V-135	110 120	125 135	1A	20A

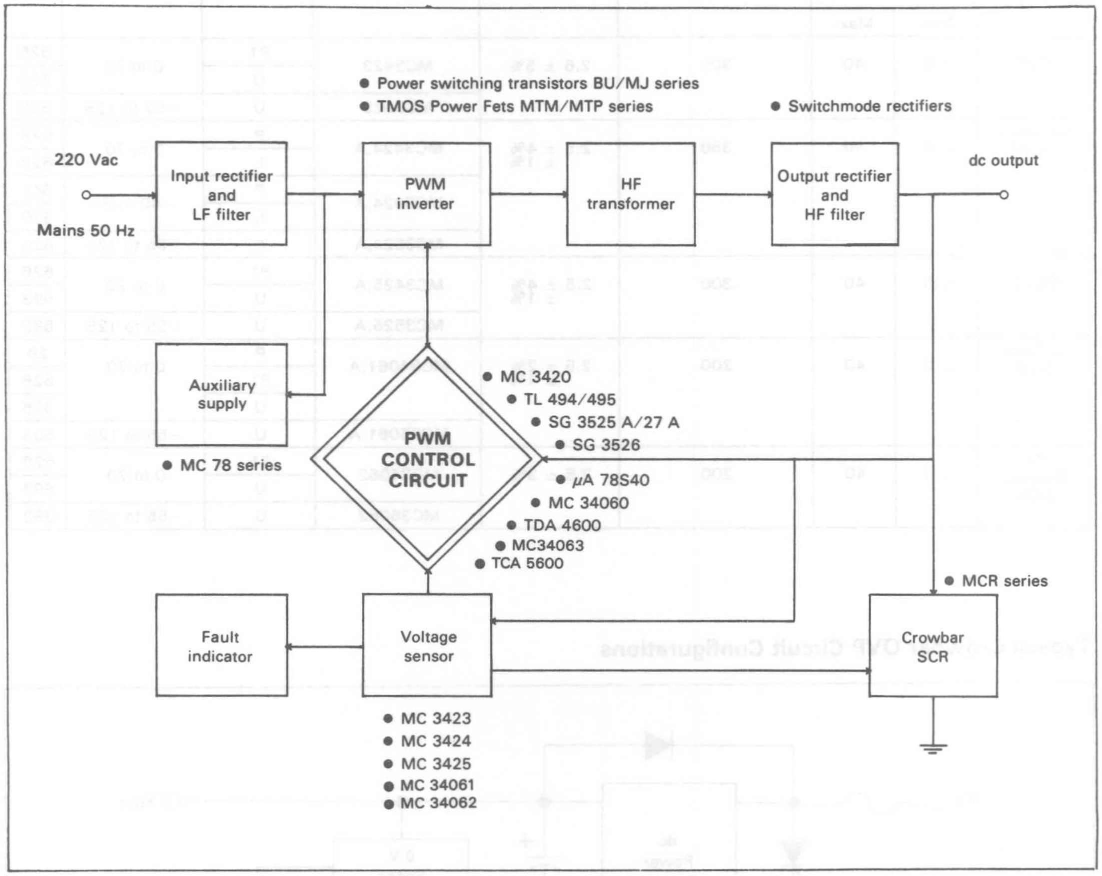


SIDAC

BILATERAL TRIGGER
Specifically designed to switch on/off directly from main in circuits as ignitor for fluorescent lamps, spark generators, flashing over voltage protection, and replacing conjonction DIAC plus TRIAC.

Integrated Circuits

What Motorola can offer for a Power Supply

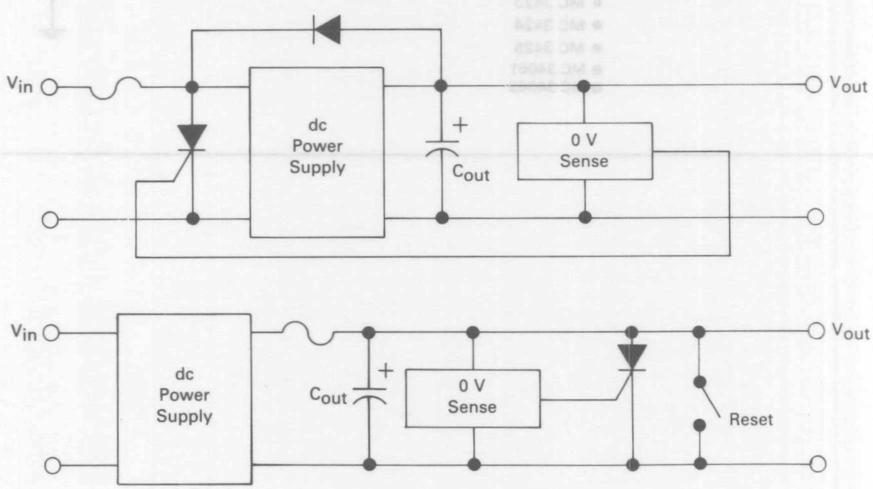


Power Supply Supervisory Circuits.

Selector Guide

TYPE	Operating Voltage Range (V) Min.	Max.	Typ. Drive Output Current (mA)	Sense trip Voltage (V)	Device Number	Suffix	Ta °C	Case
OVP	4.5	40	300	2.6 ± 5%	MC3423	P1	0 to 70	626
						U		693
					MC3523	U	− 55 to 125	693
Universal OUVP	4.5	40	350	2.5 ± 4% ± 1%	MC3424,A	P	0 to 70	648
						L		620
					MC3324,A	P	− 40 to 85	648
						L		620
					MC3524,A	L	− 55 to 125	620
OUVP	4.5	40	300	2.5 ± 4% ± 1%	MC3425,A	P1	0 to 70	626
						U		693
					MC3525,A	U	− 55 to 125	693
3 Term OVP	3.0	40	200	2.5 ± 2% ± 1%	MC34061,A	P	0 to 70	29
						P1		626
						U		626
					MC35061,A	U	− 55 to 125	693
Pin Program OVP	3.0	40	200	2.5 ± 3%	MC34062	P1	0 to 70	626
						U		693
					MC35062	U	− 55 to 125	693

Typical Crowbar OVP Circuit Configurations



MC3423
MC3523

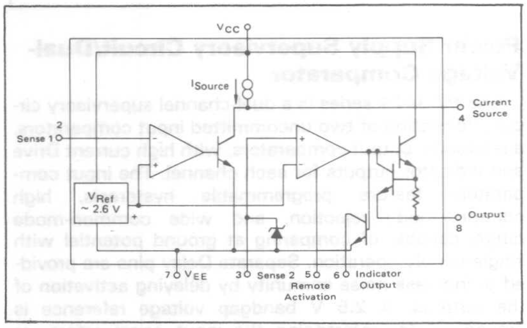
Overvoltage «Crowbar» sensing Circuit

These overvoltage protection circuits (OVP) protect sensitive electronic circuitry from overvoltage transients or regulator failures when used in conjunction with an external «crowbar» SCR. They sense the overvoltage condition and quickly «crowbar» or short circuit the supply, forcing the supply into current limiting or opening the fuse or circuit breaker.

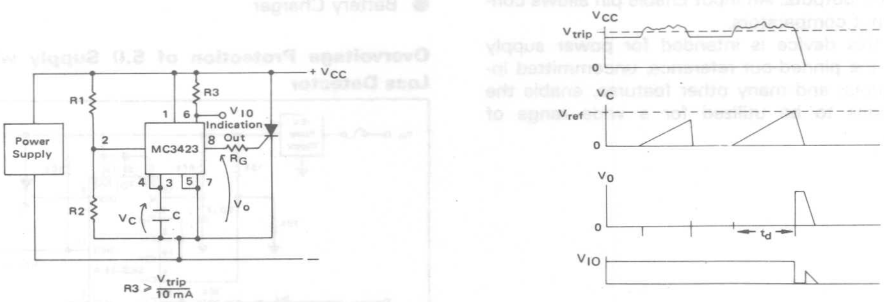
The protection voltage thershold is adjustable and the MC3423/3523 can be programmed for minimum duration of overvoltage condition before tripping, thus supplying noise immunity.

The MC3423/3523 is essentially a «two terminal» system, therefore it can be used with either positive or negative supplies.

Block Diagram



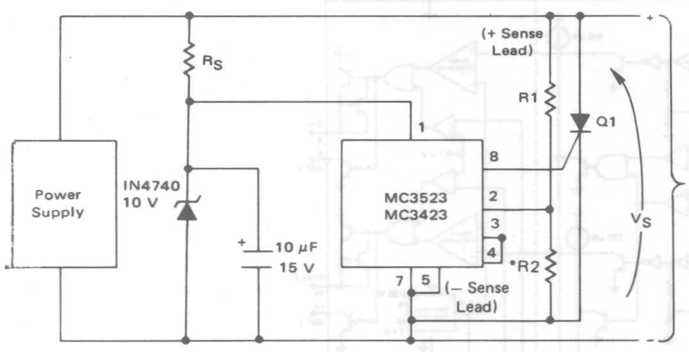
Basic Configuration for Programmable Duration of Overvoltage Condition before Trip



$R3 > \frac{V_{trip}}{10\ mA}$

$t_d = \frac{V_{ref}}{I_{source}} \times C \approx (12 \times 10^3)\ C$

Circuit Configuration for Supply Voltage above 36 V



$R_S = \left(\frac{V_S - 10}{25}\right)\ k\Omega$

$V_{trip} = V_{ref}\left(1+\frac{R1}{R2}\right) \approx 2.6\ V\left(1+\frac{R1}{R2}\right)$

$^*R2 \leqslant 10\ k\Omega$

Q1: $V_S \leqslant 50$ V; 2N6504 or equivalent
$V_S \leqslant 100$ V; 2N6505 or equivalent
$V_S \leqslant 200$ V; 2N6506 or equivalent
$V_S \leqslant 400$ V; 2N6507 or equivalent
$V_S \leqslant 600$ V; 2N6508 or equivalent
$V_S \leqslant 800$ V; 2N6509 or equivalent

MC3424, MC3424A
MC3524, MC3524A
MC3324, MC3324A

Power Supply Supervisory Circuit/Dual-Voltage Comparator

The MC3424 series is a dual-channel supervisory circuit, consisting of two uncommitted input comparators, a reference, output comparators, with high current Drive and Indicator outputs for each channel. The input comparators feature programmable hysteresis, high common-mode rejection, and wide common-mode range, capable of comparing at ground potential with single-supply operation. Separate Delay pins are provided to increase noise immunity by delaying activation of the outputs. A 2.5 V bandgap voltage reference is pinned-out for referencing the input comparators, or other external functions. Independent high current Drive and Indicator outputs for each chanel can source and sink up to 300 mA and 30 mA respectively. CMOS/TTL compatible digital inputs provide Remote Activation of each channel's outputs. An Input Enable pin allows control of the input comparators.

Although this device is intended for power supply supervision, the pinned-out reference, uncommitted input comparators, and many other features, enable the MC3424 series to be utilized for a wide range of applications.

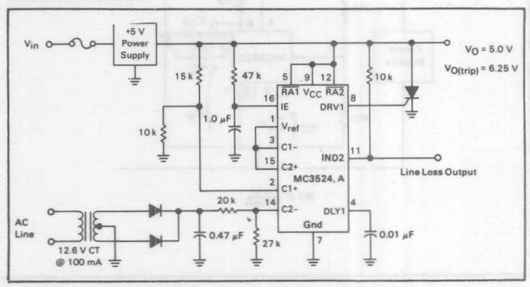
Features

- Pinned-Out 2.5 V Reference
- Wide Common-Mode Range
- Programmable Hysteresis
- Programmable Time Delays
- Two 300 mA Drive Outputs
- Remote Activation capability
- Wide Supply Range: 4.5 V ≤ V_{CC} ≤ 40 V

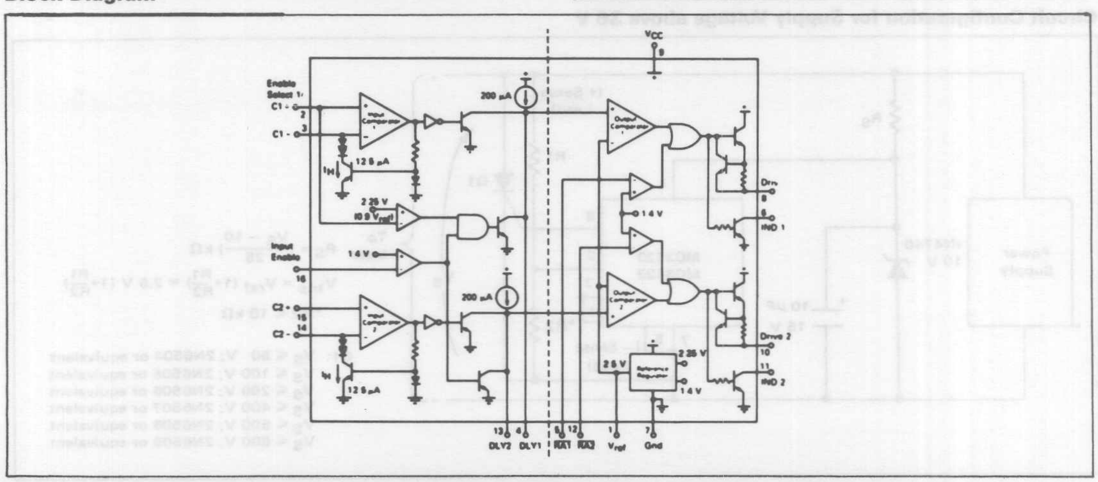
Applications

- Dual Over-Voltage «Crowbar» Protection
- Dual Under-Voltage Supervision
- Over/Under Voltage Protection
- Split-Supply Supervision
- Line-Loss Sensing
- Proportional Controller
- Programmable Frequency Switch
- Battery Charger

Overvoltage Protection of 5.0 Supply with Line Loss Detector



Block Diagram



MC3425 MC3425A
MC3525 MC3525A

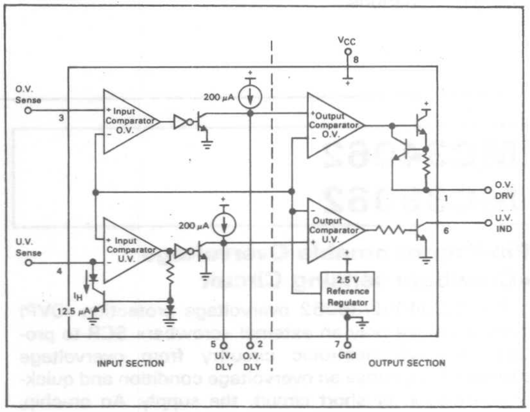
Power Supply Supervisory/over-under-Voltage Protection Circuit

The MC3425/3525 is a power supply supervisory circuit containing all the necessary functions required to monitor over- and under-voltage fault conditions. These integrated circuits contain dedicated over- and under-voltage sensing channels with independently programmable time delays. The over-voltage channels has a high current Drive Output for use in conjunction with an external SCR «Crowbar» for shutdown. The under-voltage channel input comparator has an hysteresis which is externally programmable, and an open-collector output for fault indication.

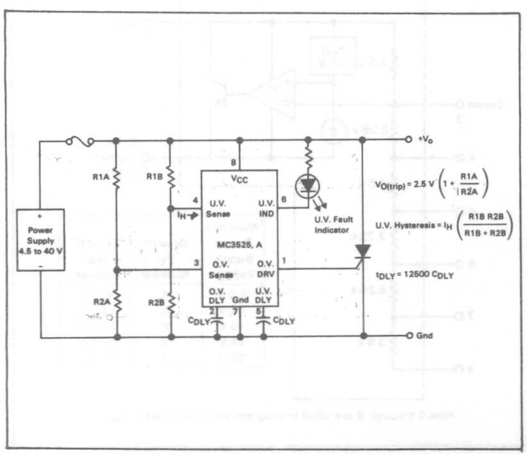
Features

- Dedicated Over- And Under-voltage Sensing
- Programmable Hysteresis Of Under-Voltage Comparator
- Internal 2.5 V Reference
- 300 mA Over-Voltage Drive Output
- 30mA Under-Voltage Indicator Output
- Programmable Time Delays
- 4.5 V to 40 V Operation

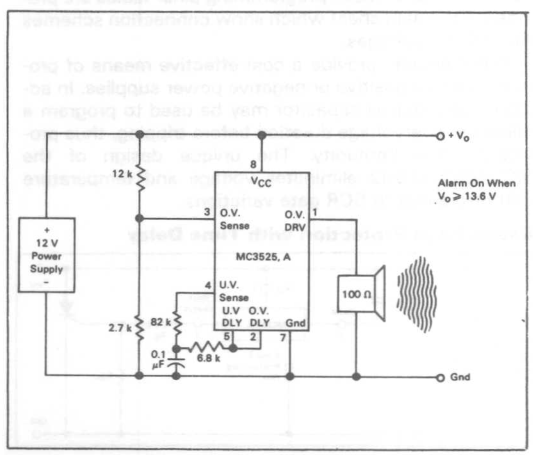
Block Diagram



Overvoltage Protection and under Voltage fault Indication with Programmable Delay



Overvoltage Audioalarm Circuit



MC34061, MC34061A
MC35061, MC35061A

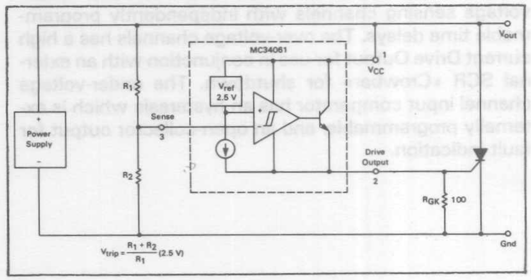
Three-Terminal Overvoltage «Crowbar» sensing Circuit

The MC34061/35061 overvoltage protection (OVP) circuits, in combination with two external programming resistors and a «crowbar» SCR, protect sensitive electronic circuitry from overvoltage damage. They sense an overvoltage condition and quickly «crowbar», or short circuit, the supply. An external capacitor may be used to program a minimum overvoltage duration before tripping, thus providing noise immunity.

These three-terminal circuits provide a cost-effective means of protecting either positive or negative power supplies. The unique design of the MC34061/35061 eliminates trip voltage and temperature drift errors due to SCR gate variations.

Features

- Unique Three-Terminal Design
- SCR Gate Drive Output of 200 mA
- Sense Voltage of 2.5 V ± 1% or ± 2%
- Hysteresis of 250 mV
- Wide Supply Range: 4.0 V ≤ V_{CC} ≤ 41 V



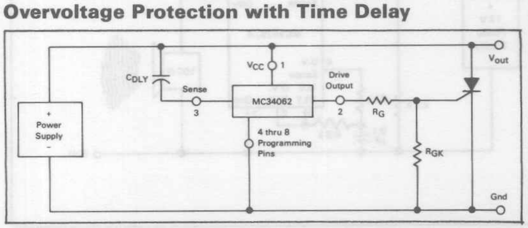
MC34062
MC35062

Pin-Programmable Overvoltage «Crowbar» sensing Circuit

The MC34062/35062 overvoltage protection (OVP) circuits require only an external «crowbar» SCR to protect sensitive electronic circuitry from overvoltage damage. They sense an overvoltage condition and quickly «crowbar», or short circuit, the supply. An on-chip, tapped resistor network allows the device to be programmed for trip voltages ranging from 3.5 to 40 V. Each of the five programming pins provides one standard overvoltage trip point for nominal power supply voltages of 5.0, 12, 15, 24 or 28 V. Many other trip voltages may be programmed by interconnecting and grounding various combinations of these programming pins. Tables are provided in the data sheet which show connection schemes for 120 trip voltages.

These circuits provide a cost-effective means of protecting either positive or negative power supplies. In addition, an external capacitor may be used to program a minimum overvoltage duration before tripping, thus providing noise immunity. The unique design of the MC34062/35062 eliminates voltage and temperature drift errors due to SCR gate variations.

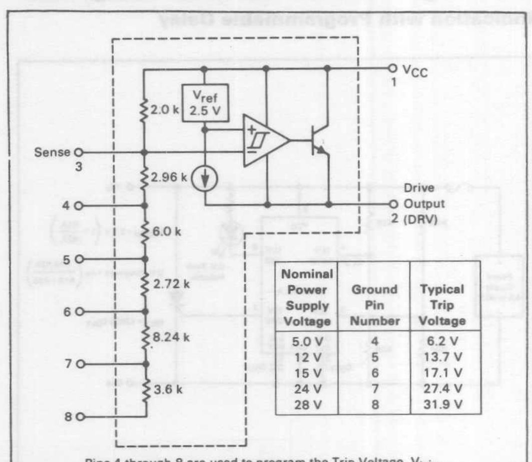
Overvoltage Protection with Time Delay



Features

- Unique Pin-Programmable Trip Voltage from 3.5 to 40 V
- One-Pin Programming for 5.0, 12, 15, 24 and 28 V Power Supplies
- SCR Gate Drive Output of 200 mA
- Built-In Hysteresis Voltage
- Wide Supply Range: 4.0 V ≤ V_{CC} ≤ 41 V

Functional Block Diagram

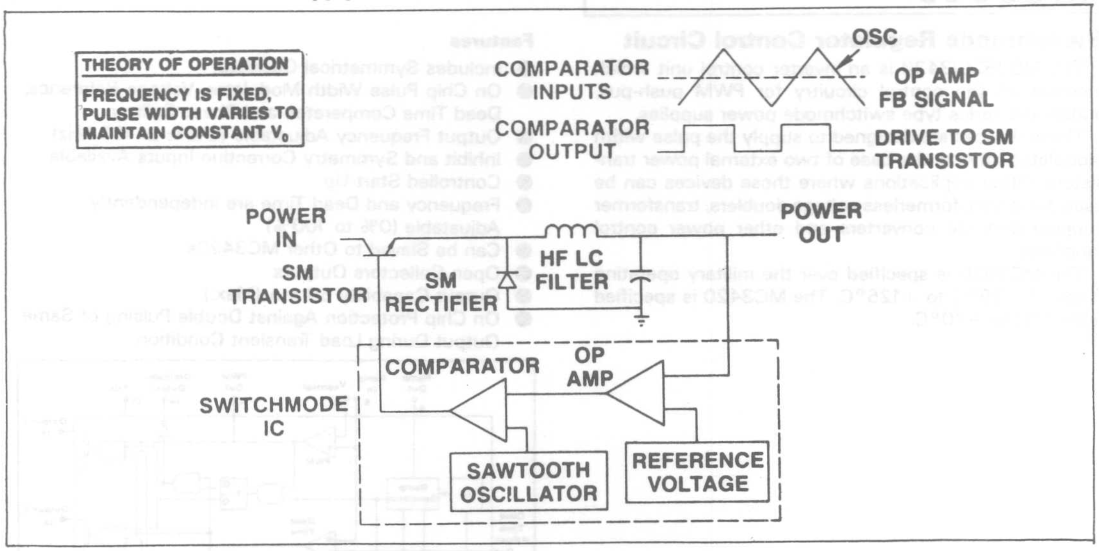


Nominal Power Supply Voltage	Ground Pin Number	Typical Trip Voltage
5.0 V	4	6.2 V
12 V	5	13.7 V
15 V	6	17.1 V
24 V	7	27.4 V
28 V	8	31.9 V

Pins 4 through 8 are used to program the Trip Voltage, V_{trip}

Switchmode Regulator Control Circuits.

What is Switchmode Power Supply?



Pulse Width Modulation Control IC's Selector Guide

Io mA Max	Vcc (V) Min	Vcc (V) Max	fo (kHz) Min	fo (kHz) Max	Device number	Suffix	Temperature range TA (°C)	Case
50 100*	10	30	2.0	100	MC 3420	P	0 to + 70	648
						L		620
					MC 3520	L	− 55 to + 125	620
200	7	40	1.0	200	MC34060	P	0 to + 70	646
						L		632
					MC35060	L	− 55 to + 125	632
± 400	8	35	0.1	500	SG 1525 A, 27 A	J	− 55 to + 125	620
					SG 2525 A, 27 A	N	− 40 to + 85	648
						J		620
					SG 3525 A, 27 A	N	0 to + 70	648
						J		620
± 200	8	35	0.001	400	SG 1526	J	− 55 to + 125	726
					SG 2526	N	− 40 to + 85	707
						J		726
					SG 3526	N	0 to + 70	707
						J		726
250 500*	7	40	1.0	300	TL 494	CN	0 to + 70	648
						CJ		620
						IN	− 40 to + 85	648
						IJ		620
						MJ	− 55 to + 125	620
		> 40**			TL495	CN	0 to + 70	701
						CJ		726
						IN	− 40 to + 85	701
						IJ		726
						MJ	− 55 to + 125	726
1500	2.5	40	1	40	µA 78S40	DM	− 55 to 125	620
						DC	0 to + 70	620
						PC		648
	2.5	40	0.1	100	MC34063	P1	0 to + 70	626
						U		693
					MC33063	P1	− 40 to + 85	626
						U		693
					MC35063	U	− 55 to + 125	693
	8	20	20	70	TDA4600	—	− 15 to + 85	762

* In single ended configuration.
** TL 495 features a 39 V zener for high voltage operation.

MC3420
MC3520

Switchmode Regulator Control Circuit

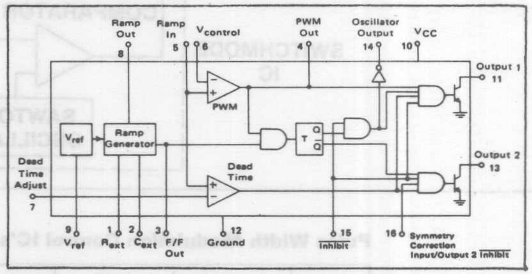
The MC3520/3420 is an inverter control unit which provides all the control circuitry for PWM push-pull, bridge and series type switchmode power supplies.

These devices are designed to supply the pulse width modulated drive to the base of two external power transistors. Other applications where these devices can be used are in transformerless voltage doublers, transformer coupled dc-to-dc converters and other power control functions.

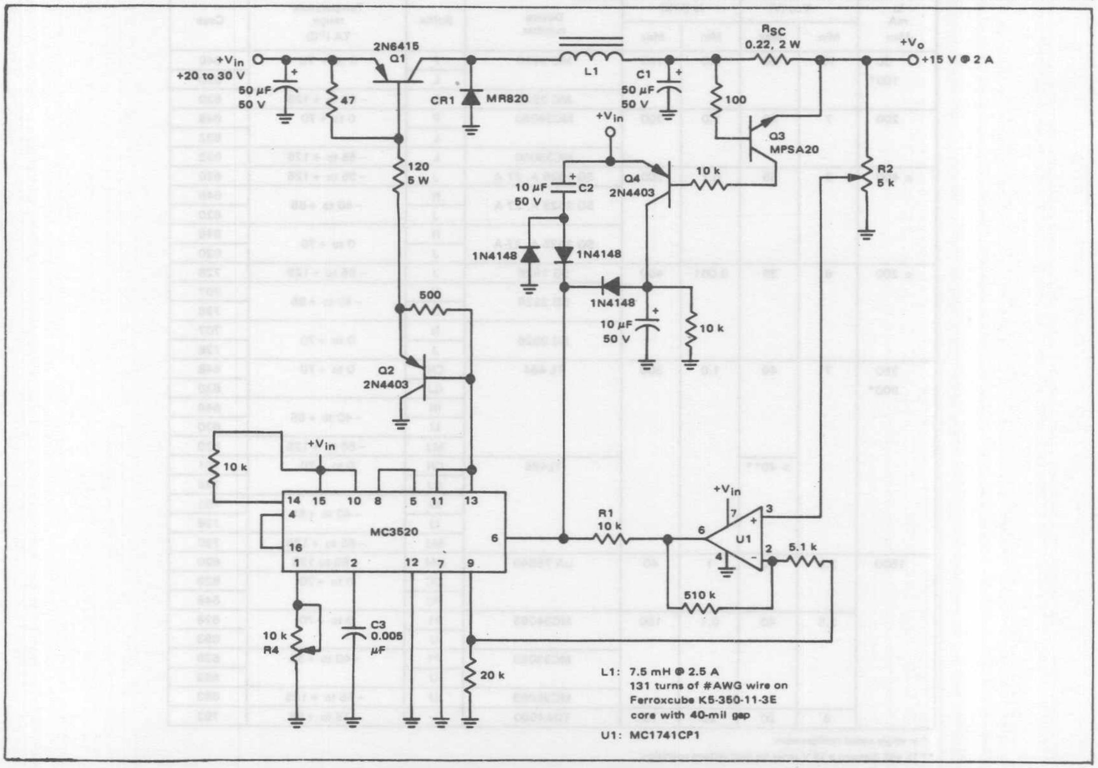
The MC3520 is specified over the military operating range of −55°C to +125°C. The MC3420 is specified from 0°C to +70°C.

Features

- Includes Symmetrical Oscillator
- On Chip Pulse Width Modulator, Voltage Reference, Dead Time Comparator, and Phase Splitter
- Output Frequency Adjustable (2 kHz to 100 kHz)
- Inhibit and Symmetry Correction Inputs Available
- Controlled Start-Up
- Frequency and Dead Time are Independently Adjustable (0% to 100%)
- Can be Slaved to Other MC3420s
- Open Collectors Outputs
- Output Capability 50 mA (Max.)
- On Chip Protection Against Double Pulsing of Same Output During Load Transient Condition



15V, 2A DC to DC Converter



TL494
TL495

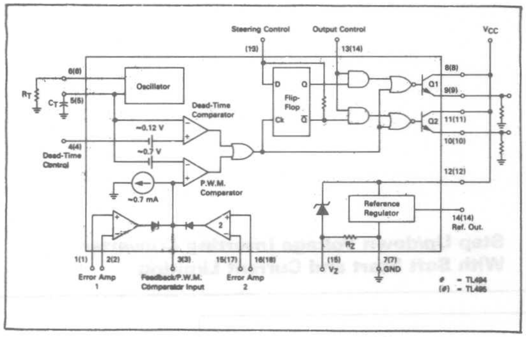
Switchmode Pulse Width Modulation Control Circuits

The TL494 and TL495 are fixed frequency, pulse width modulation control circuits designed primarily for Switchmode power supply control. These devices feature:

Features

- Complete Pulse Width Modulation Control Circuitry
- On-Chip Oscillator With Master Or Slave Operation
- On-Chip Error Amplifiers
- On-Chip 5 Volt Reference
- Adjustable Dead-Time Control
- Uncommitted Output Transistors For 200 mA Source Or Sink
- Output Control For Push-Pull Or Single-Ended Operation
- On-Chip 39 Volt Zener (TL495 Only)
- Output Steering Control (TL495 Only)

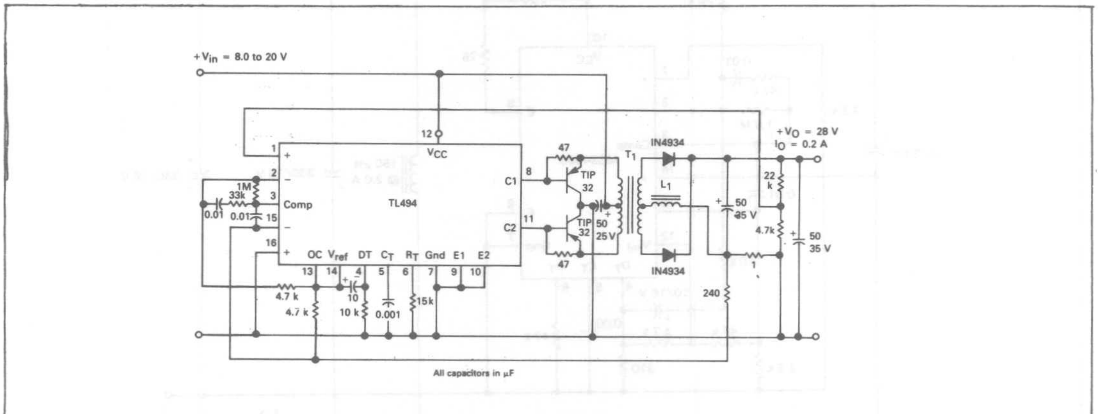
Block Diagram



Pulse Width Modulated Push-Pull Converter

The TL 494 control circuit is used in the push-pull mode to drive power transistors connected to the high frequency transformer T1. The voltage developed at the secondary winding is rectified and smoothed by the filter components.

Typical Application Configuration



TEST	CONDITIONS	RESULTS
Line Regulation	V_{in} = 10V to 40V	14mV 0.28%
Load Regulation	V_{in} = 28V, I_O = 1 mA to 1 A	3.0mV 0.06%
Output Ripple	V_{in} = 28V, I_O = 1.0 A	65mV P-P P.A.R.D.
Short Circuit Current	V_{in} = 28V, R_L = 0.1Ω	1.6 amps
Efficiency	V_{in} = 28V, I_O = 1A	71%

MC34060
MC35060

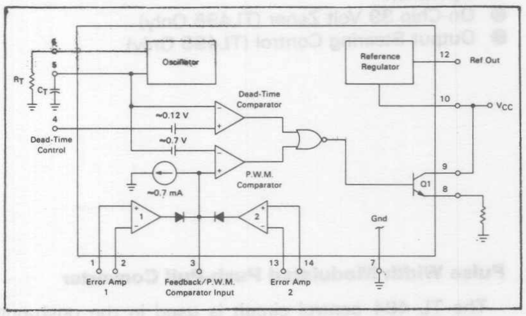
Switchmode Pulse width Modulation Control Circuits

The MC35060 and MC34060 are low cost fixed frequency, pulse width modulation control circuits designed primarily for single ended switchmode power supply control. These devices feature:

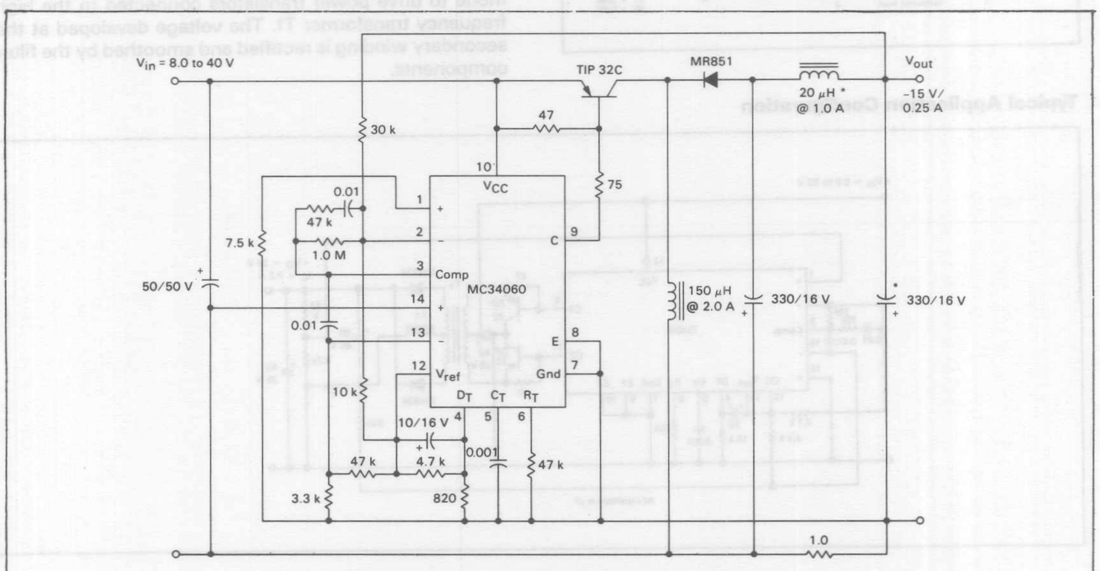
Features

- Complete Pulse Width Modulation Control Circuitry
- On-Chip Oscillator With Master or Slave Operation
- On-Chip Error Amplifiers
- On-Chip 5.0 Volt Reference
- Adjustable Dead Time Control
- Uncommitted Output Transistor for 200 mA Source or Sink

Block Diagram



Step Up/down Voltage Inverting Converter With Soft Start and Current Limiting



TEST	CONDITIONS	RESULTS
Line Regulation	V_{in} = 80V to 40V, I_O = 250 mA	52mV 0.35%
Load Regulation	V_{in} = 12V, I_O = 1 mA to 250 mA	47mV 0.32%
Output Ripple	V_{in} = 12V, I_O = 250 mA	10mV p-p P.A.R.D.
Short Circuit Current	V_{in} = 12V, R_L = 0.1Ω	330 mA
Efficiency	V_{in} = 12V, I_O = 250 mA	86%

SG1525A/SG1527A
SG2525A/SG2527A
SG3525A/SG3527A

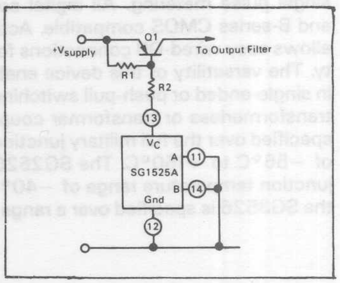
Pulse width Modulation Control Circuit

The SG1525A/1527A series of pulse width modulator control-circuits offer improved performance and lower external parts count when implemented for controlling all types of switching power supplies. The device includes a +5.1 volt ±1% reference and an error amplifier with a common-mode range including the reference voltage to eliminate external divider resistors. A sync input to the oscillator enables multiple units to be slaved together, or a single unit can be synchronized to an external system clock. A wide range of dead time is programmable with a single resistor between the C_T pin and the Discharge pin. Other features included are soft-start circuitry requiring only an external timing capacitor. A shutdown pin controls both the soft-start circuitry and the output stages, allowing fast output turn-off with soft-start recycle turn-on. Undervoltage lockout keeps the outputs off when V_{CC} is less than the required level for normal operation. The output stages are a totem-pole design capable of sinking and sourcing in excess of 200 mA. The SG1525A series output stage features NOR Logic, giving a low output for an off state. The SG1527A utilizes OR Logic which results in a high output level when off. These devices are available in Military, Industrial and Commercial temperature ranges.

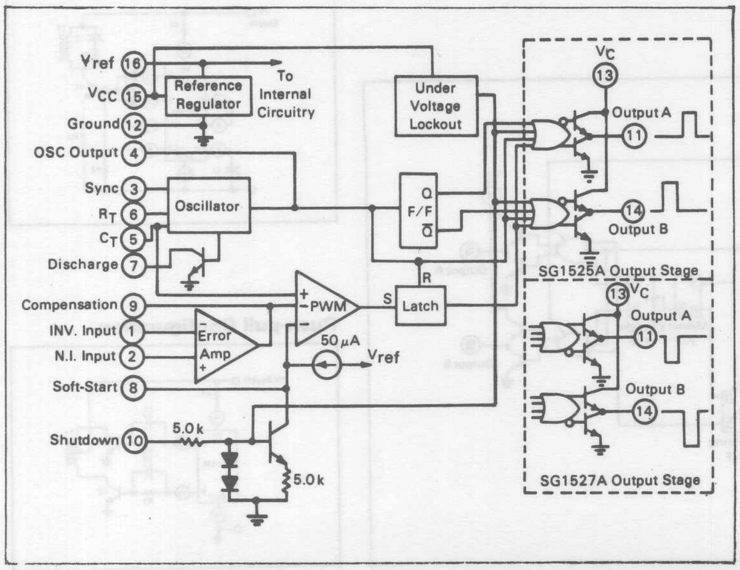
Features

- 8.0 to 35 Volt Operation
- 5.1 Volt ±1% Trimmed Reference
- 100 Hz to 400 kHz Oscillator Range
- Separate Oscillator Sync Pin
- Adjustable Dead Time
- Input Undervoltage Lockout
- Latching PWM to Prevent Multiple Pulses
- Dual Source/Sink Output Current: ±400 mA Peak

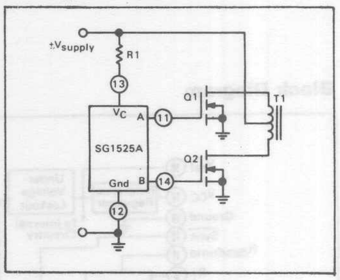
Single ended Supply



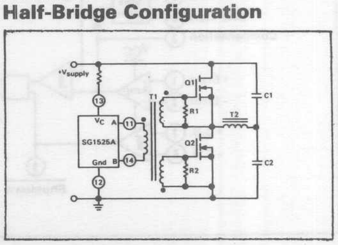
Bloc Diagram



Driving Power Fets



Driving Transformers in a Half-Bridge Configuration



SG1526
SG2526
SG3526

Pulse width Modulation Control Circuit

The SG1526 is a high performance pulse width modulator integrated circuit intended for fixed frequency switching regulators and other power control applications.

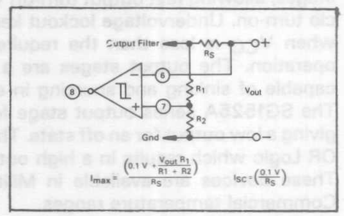
Functions included in this IC are a temperature compensated voltage reference, sawtooth oscillator, error amplifier, pulse width modulator, pulse metering and steering logic, and two high current totem pole outputs ideally suited for driving the capacitance of power FETs at high speeds.

Additional protective features include soft-start and undervoltage lockout, digital current limiting, double pulse inhibit, adjustable dead time and a data latch for single pulse metering. All digital control ports are TTL and B-series CMOS compatible. Active low logic design allows easy wired-OR connections for maximum flexibility. The versatility of this device enables implementation in single-ended or push-pull switching regulators that are transformerless or transformer coupled. The SG1526 is specified over the full military junction temperature range of −55°C to +150°C. The SG2526 is specified over a junction temperature range of −40°C to +150°C while the SG3526 is specified over a range of 0°C to +125°C.

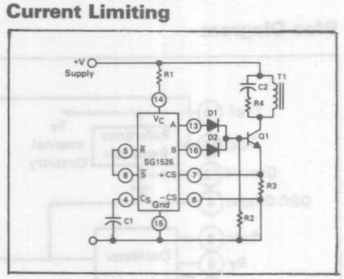
Features

- 8.0 to 35 Volt Operation
- 5.0 Volt ±1% Trimmed Reference
- 1.0 Hz to 400 kHz Oscillator Range
- Dual Source/Sink Current Outputs: ±100 mA
- Digital Current Limiting
- Programmable Dead Time
- Undervoltage Lockout
- Single Pulse Metering
- Programmable Soft-Start
- Wide Current Limit Common Mode Range
- Guaranteed 6 Unit Synchronization

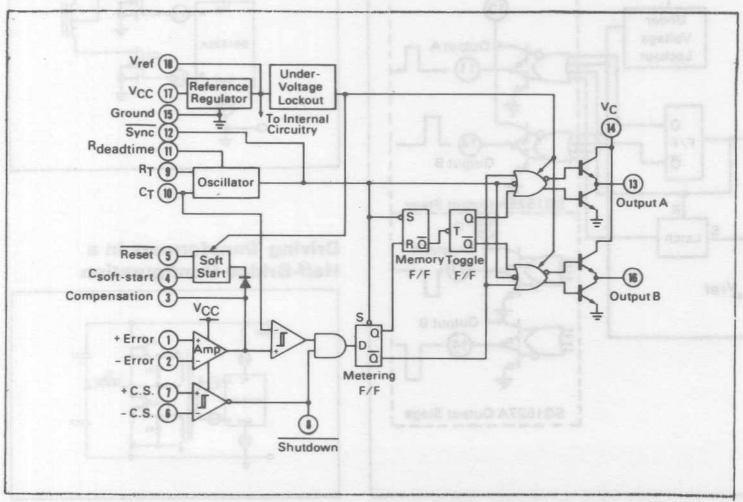
Foldback Current Limiting



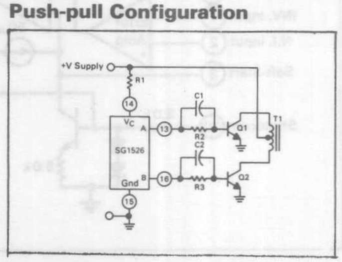
Flyback Converter with Current Limiting



Block Diagram



Push-pull Configuration



μA78S40

Universal Switching Regulator Subsystem

The μA78S40 is a monolithic-switching regulator subsystem, providing all active functions necessary for a switching regulator system. The device consists of a tight-tolerance temperature-compensated voltage reference, controlled-duty cycle oscillator with an active peak-current limit circuit, comparator, high-current and high-voltage output switch, capable of 1.5 A and 40 V, pinned-out power diode and an uncommitted operational amplifier, which can be powered up or down independent of the I.C. supply. The switching output can drive external NPN or PNP transistors when voltages greater than 40 V, or currents in excess of 1.5 A, are required. Some of the features are wide-supply voltage range, low standby current, high efficiency and low drift. The μA78S40 is available in both commercial (0°C to +70°C) and military (−55°C to +125°C) temperature ranges.

Some of the applications include use in step-up, step-down, and inverting regulators, with extremely good results obtained in battery-operated systems.

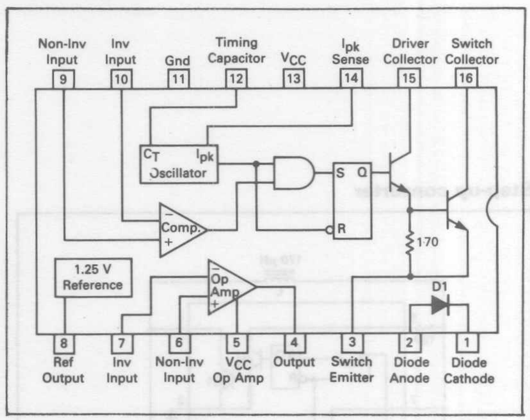
Step-Down Voltage Regulator

This application illustrates the necessary connections for using the μA78S40 in the step-down mode. This version satisfies a requirement for a 5 V output at 500 mA with output ripple less than 25 mV. In this application, the power switch (Q1 and Q2) is connected between the +25 V supply and the inductor. Switch voltage is determinated by the emitter output limit of 1.6 V. For this value, off-time is approximately three times the on-time. Considering that the on-time should be greater than 10 μs, off-time is set by the user at 60 μs (C_T = 0.02 μH), and the inductor value is selected at 330 μF. The output voltage is set by resistors R1 and R2. Output capacitance is calculated to be 400 μF. The circuit standby power is less than 50 mW. Efficiency at full load is 79%; at 10% of full load it is 70%.

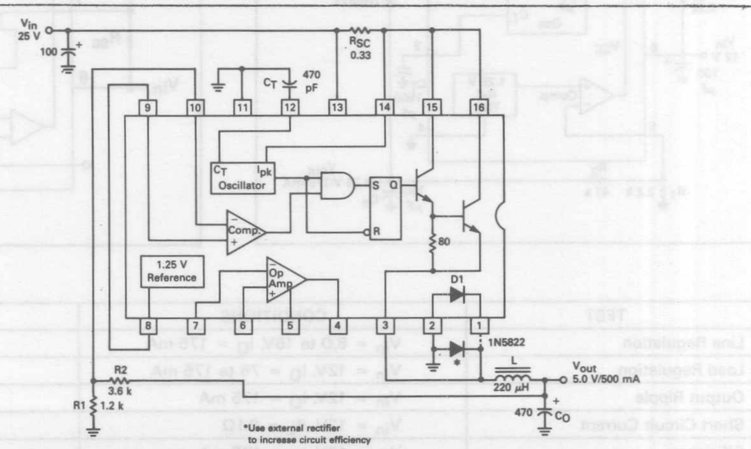
Features

- Output Adjustable from 1.3 V to 40 V
- Peak Output Current of 1.5 A Without External Transistor
- 80 dB Line and Load Regulation
- Operation from 2.5 V to 40 V Supply
- Low Standby Current Drain
- High Gain, High Output Current, Uncommitted Op Amp.
- Uncommitted Power Diode
- Low Cost

Block Diagram



Step-down Converter



MC34063
MC35063
MC33063

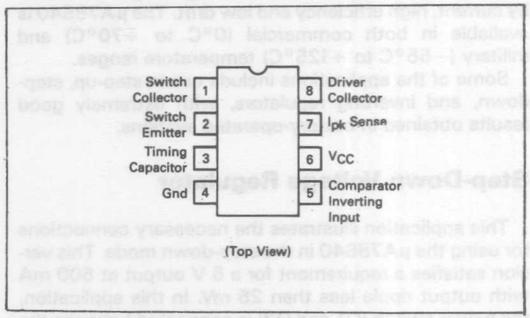
DC to DC Converter Control Circuits

The MC34063 Series is a monolithic control circuit containing the primary functions required for dc-to-dc converters. The device consists of an internal temperature compensated reference, comparator, controlled duty cycle oscillator with an active current limit circuit, driver and high current output switch. This series was specifically designed to be incorporated in Step-Down (Buck) and Step-Up (Boost) applications with a minimum number of external components.

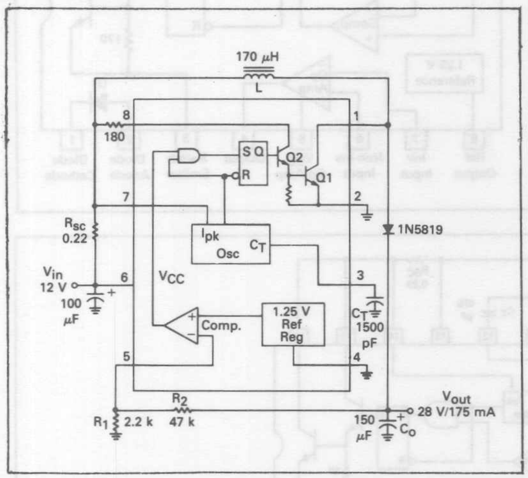
Features

- Operation from 2.5 V to 40 V Input
- Low Standby Current
- Current Limiting
- Output Switch Current of 1.5 A
- Output Voltage Adjustable from 1.25 to 40 V
- Frequency Operation from 100 Hz to 100 kHz

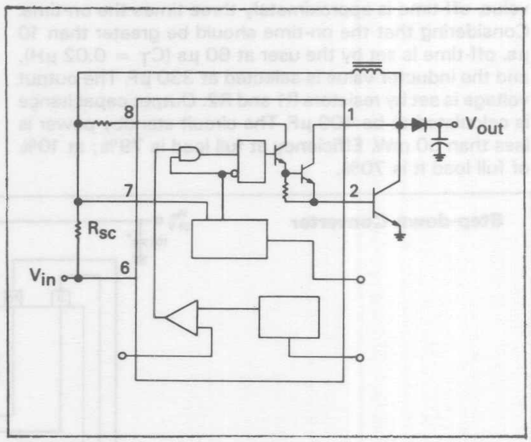
Pin Connections



Step-up converter



External NPN Switch



TEST	CONDITIONS	RESULTS
Line Regulation	V_{in} = 8,0 to 16V, I_O = 175 mA	12mV
Load Regulation	V_{in} = 12V, I_O = 75 to 175 mA	45mV
Output Ripple	V_{in} = 12V, I_O = 175 mA	150 mV_{p-p}
Short Circuit Current	V_{in} = 12V, R_L = 0.1Ω	2.0 A
Efficiency	V_{in} = 12V, I_O = 175 mA	93%

TCA5600
TCF5600

Universal Microprocessor Power Supply Controller

The TCA5600 is a versatile power supply control circuit for microprocessor based systems and mainly intended for automotive applications and battery powered instruments. To cover a wide range of applications, the device offers high circuit flexibility with minimum of external components.

Functions included in this IC are a temperature compensated voltage reference, on chip dc/dc converter, programmable and remote controlled voltage regulator, fixed 5.0 V supply voltage regulator with external PNP power device, undervoltage detection circuit, power-on RESET delay and watchdog feature for safe and hazard free microprocessor operations.

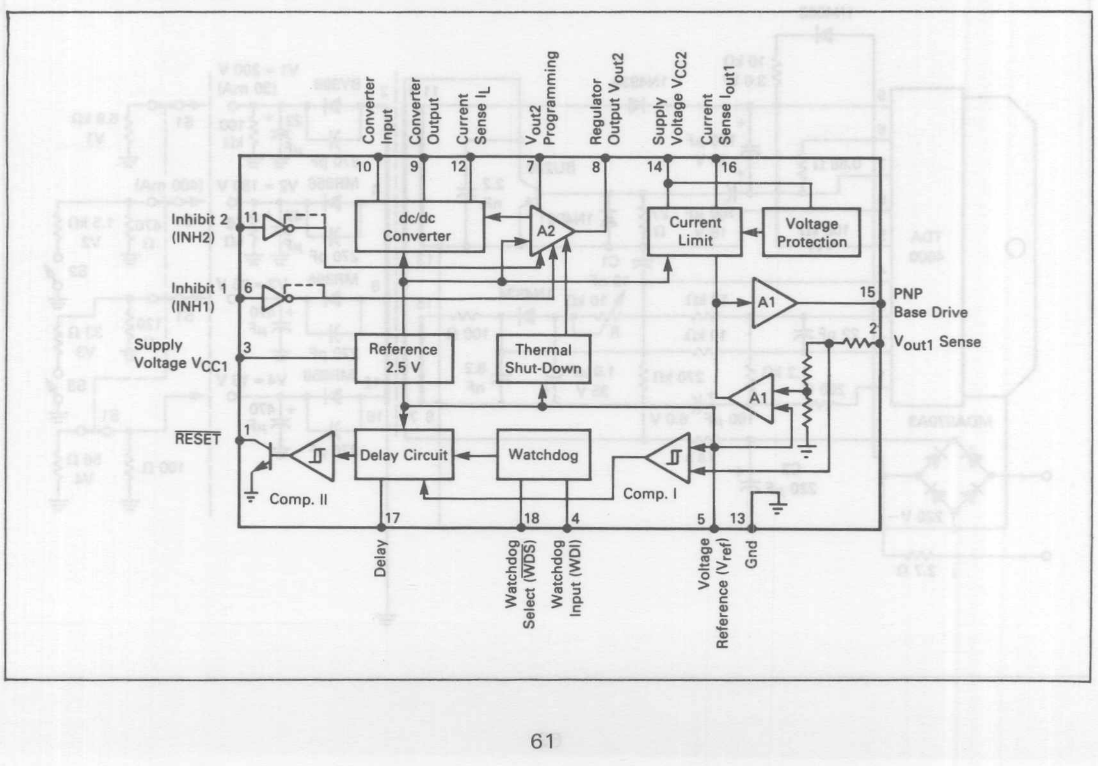
Features

- 6.0 to 30 V Operation Range
- 2.5 V Reference Voltage Accessible for Other Tasks
- Fixed 5.0 V ± 4% Microprocessor Supply Regulator Including Current Limitation, Overvoltage Protection and Undervoltage Monitor
- Programmable 6.0 to 30 V Voltage Regulator Exhibiting High Peak Current (150 mA), Current Limiting and Thermal Protection
- Two Remote Inputs to Select the Regulator's Operation Mode: OFF, 5.0 V, 5.0 V Standby and Programmable Output Voltage
- Self Contained dc/dc Converter Fully Controlled By the Programmable Regulator to Guarantee Safe Operation Under All Working Conditions
- Programmable Power-On RESET Delay
- Watchdog Select Input
- Negative Edge Triggered Watchdog Input
- Low Current Consumption in the V_{CC1} Standby Mode
- All Digital Control Ports are TTL- and MOS-Compatible

APPLICATIONS INCLUDE

- Microprocessor Systems with E^2PROMs
- High Voltage Crystal and Plasma Displays
- Decentralized Power Supplies in Computer and Telecommunication Systems

Functional Block Diagram



TDA4600

Control IC for Mains Isolated freely Oscillating Flyback Converter

The bipolar integrated circuit TDA4600 drives, regulates and monitors the switching transistor in a power supply based on freely oscillating converters.

Due to the wide regulating range and the high voltage stability during large load changes, SMPS for Hi-Fi equipment and active loudspeakers can be realized as well as applications in TV receivers and video recorders.

The TDA4600 is available in a 9-pins SIP plastic medium power package. The ambient temperature during operation can be from −15°C to +85°C.

This application represents a blocking converter for color TV sets with 30 W to 120 W of output power ans mains voltages from 160 V to 270 V.

This circuit shows the low number of external components. Inspite of regulation on the primary side, high voltage stability of the various secondary voltages is achieved even with large load changes.

For mains isolation and transformation to the desired secondary voltages, a transformer with ferrite core is used.

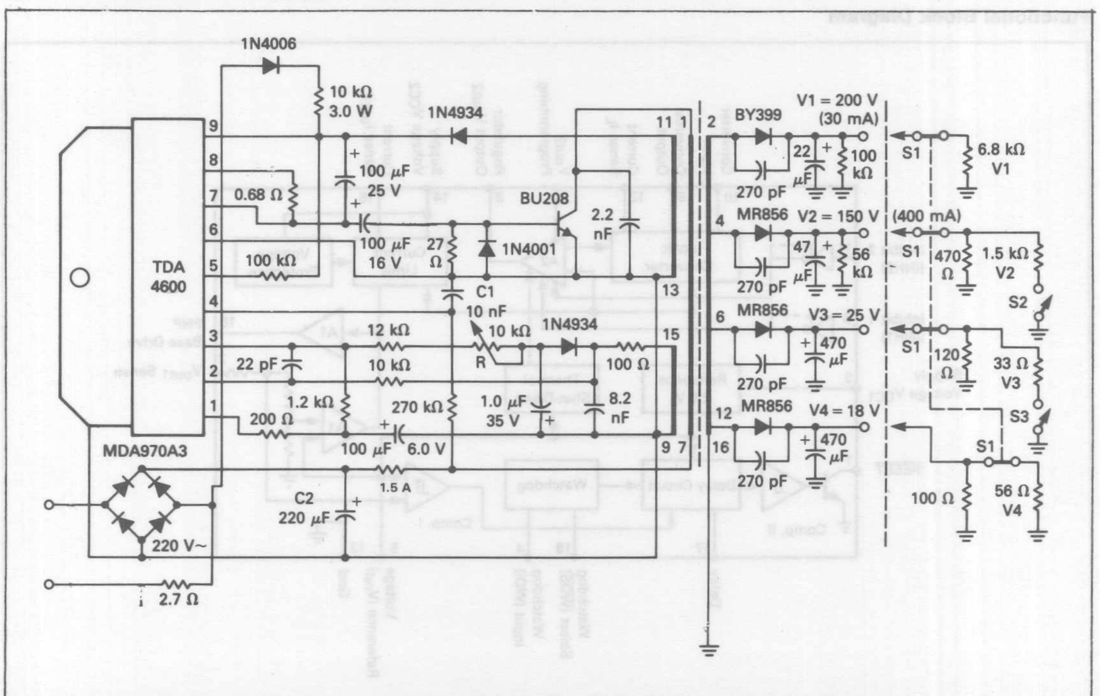
Features

- Direct driving of the power switching transistor
- Low starting current, defined starting behaviour also at slowly rising mains voltage.
- Short-circuit proof and open-loop resistant circuit. In both cases a power of only 6 to 10 W is consumed. Linear foldback characteristic at overload
- Automatic restart after elimination of the overload
- Efficiency of more than 80% at an output power of 40 to 100 W
- Frequency of oscillation between 20 kHz (100 W) and 70 kHz (without load)
- Simple RF I suppression
- Good regulation of load current and mains voltage variations. At a mains voltage variation between 170 and 240 V the output voltage of 150 V will change only by about 2 V.

Features

- Wide operational range
- High voltage stability even at high load changes
- Direct control of switching transistor
- Low start-up current
- Linear foldback of the overload characteristic
- Base drive proportional to the current through the power switching transistor

Flyback Converter Power Supply



Motor Control Circuits

Device	Icc	Iout	Soft Start	Firing Quadrant	Features	Package	Pins	Case
TDA 1085C	4,2mA	150mA	X	1-4	Motor current limiting	Plastic	16	648
TDA 1185A	1,0mA	min80mA	X	2-3	Firing pulse repetition if triac fails	Plastic	14	646
TDA 1285A	4,5mA	min45mA	X	1-4	Feed back with Hall effect IC	Plastic	16	648

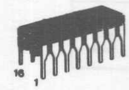
Zero Voltage Switches

Device	Icc	Iout	Vio	Features	Package	Pins	Case
UAA 1004	1,9mA	min80mA	± 10mV	Built-in Hysteresis – Sensor fail protection	Plastic	8	626
UAA 1016 B	1,5mA	min80mA typ50mA	± 5mV	Burst control – Sensor fail protection	Plastic	8	626

Stepper Motor Drivers

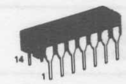
Device	Vcc	Vm	Im	Features	Package	Pins	Case
SAA1042	20V	12V	0,5A	Half/Full Step Operation Clockwise/Counter Clockwise Operation	Plastic	12	721-02
SAA1042A	20V	24V	0,5A	MOS – TTL – DTL Compatible	Plastic	12	721-02

Package Styles

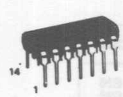


L SUFFIX
CERAMIC PACKAGE
CASE 620

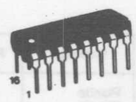
P SUFFIX
PLASTIC PACKAGE
CASE 626



L SUFFIX
CERAMIC PACKAGE
CASE 632

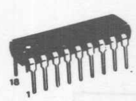


P SUFFIX
PLASTIC PACKAGE
CASE 646



P SUFFIX
PLASTIC PACKAGE
CASE 648

U SUFFIX
CERAMIC PACKAGE
CASE 693

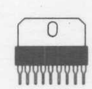


N SUFFIX
PLASTIC PACKAGE
CASE 701

P SUFFIX
PLASTIC PACKAGE
CASE 721

L SUFFIX
CERAMIC PACKAGE
CASE 726

P SUFFIX
PLASTIC PACKAGE
CASE 29
TO-92



SIP 9
PLASTIC POWER PACKAGE
CASE 762-01

Cross Reference

REFERENCE NUMBER	MOTOROLA DIRECT REPLACEMENT	MOTOROLA FUNCTIONAL EQUIVALENT
CA3524		TL494
LAS3800		TL494
LM3524		TL494
MB3759		MC3420
NE5560		TDA4600
NE5561		MC34060
PWM125	SG1525A	
RC4190		µA78S40
SI1525B	SG1525A	
SI1527B	SG1527A	
SG3523	MC3423	
SG3524		TL494
SG3542		MC3423
SG3543		MC3425
SG3544		MC3424
TDA1060		TDA4600
TDA4718		SG1526
TEA1039		TDA4600
TL497		µA78S40
TL594	TL494	
TL595	TL495	
TL7702		MC3423
TL7705		MC34062
UAA4001		TL494
UAA4006		TDA4600
UC494	TL494	
UC495	TL495	
UC1524		TL494
UC1525A	SG1525A	
UC1527A	SG1527A	
ULN8125	SG1525A	
ULN8126	SG1526	
ULN8127	SG1527A	
ULN8160		TDA4600
ULN8161		MC34060
ULN8194	TL494	
ULN8195	TL495	
µA494	TL494	
µPC3423	MC3423	
XR494	TL494	
XR495	TL495	
XR1524		TL494
XR2230		SG1526
ZN1060		TDA4600
ZN1066		TL494